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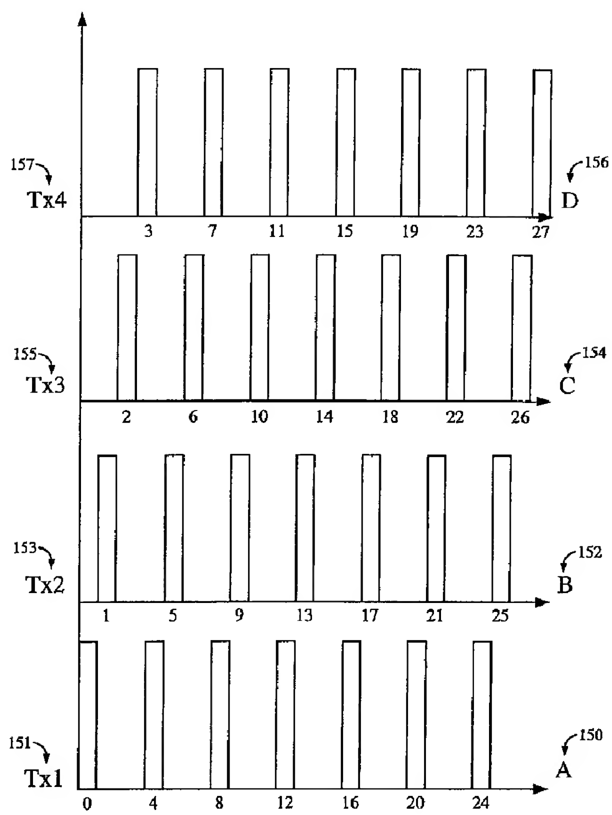
(54) Titre : PROCEDE ET APPAREIL DE MESURE ET DE SIGNALEMENT DES INFORMATIONS RELATIVES A L'ETAT DES CANAUX DANS UN SYSTEME DE COMMUNICATIONS A EFFICACITE ET PERFORMANCE ELEVEES

(54) Title: METHOD AND APPARATUS FOR MEASURING AND REPORTING CHANNEL STATE INFORMATION IN A HIGH EFFICIENCY, HIGH PERFORMANCE COMMUNICATIONS SYSTEM



OFDM SUB-CHANNELS

(57) Abrégé/Abstract:
Channel state information (CSI) can be used by a communications system to precondition transmissions between transmitter units and receiver units. In one aspect of the invention, disjoint sub-channel sets are assigned to transmit antennas located at a

(57) **Abrégé(suite)/Abstract(continued):**
transmitter unit. Pilot symbols are generated and transmitted on a subset of the disjoint sub-channels. Upon receipt of the transmitted pilot symbols, the receiver units determine the CSI for the disjoint sub-channels that carried pilot symbols. These CSI values are reported to the transmitter unit, which will use these CSI values to generate CSI estimates for the disjoint sub-channels that did not carry pilot symbols. The amount of information necessary to report CSI on the reverse link can be further minimized through compression techniques and resource allocation techniques.

Abstract

Channel state information (CSI) can be used by a communications system to precondition transmissions between transmitter units and receiver units. In one aspect of the invention, disjoint sub-channel sets are assigned to transmit antennas located at a transmitter unit. Pilot symbols are generated and transmitted on a subset of the disjoint sub-channels. Upon receipt of the transmitted pilot symbols, the receiver units determine the CSI for the disjoint sub-channels that carried pilot symbols. These CSI values are reported to the transmitter unit, which will use these CSI values to generate CSI estimates for the disjoint sub-channels that did not carry pilot symbols. The amount of information necessary to report CSI on the reverse link can be further minimized through compression techniques and resource allocation techniques.

METHOD AND APPARATUS FOR MEASURING AND REPORTING CHANNEL STATE INFORMATION IN A HIGH EFFICIENCY, HIGH PERFORMANCE COMMUNICATIONS SYSTEM

This application is a divisional of Canadian Patent Application No. 2,402,152 filed March 20, 2001.

BACKGROUND OF THE INVENTION

I. Field of the Invention

The present invention relates to the field of communications. More particularly, the present invention relates to the measurement and report of channel state information in a high efficiency, high performance communications system.

II. Description of the Related Art

A modern day wireless communications system is required to operate over channels that experience fading and multipath. One such communications system is a code division multiple access (CDMA) system that conforms to the "TIAEIA/IS-95 Mobile Station-Base Station Compatibility Standard for Dual-Mode Wideband Spread Spectrum Cellular System", hereinafter referred to as the IS-95 standard. The CDMA system supports voice and data communication between users over a terrestrial link. The use of CDMA techniques in a multiple access communication system is disclosed in U.S. Patent No. 4,901,307, entitled "SPREAD SPECTRUM MULTIPLE ACCESS COMMUNICATION SYSTEM USING SATELLITE OR TERRESTRIAL REPEATERS", and U.S. Patent No. 5,103,459, entitled "SYSTEM AND METHOD FOR GENERATING WAVEFORMS IN A CDMA CELLULAR TELEPHONE SYSTEM", both assigned to the assignee of the present invention and incorporated herein by reference.

An IS-95 system can operate efficiently by estimating channel parameters at a receiver unit, which uses these estimated channel parameters to

from every base station. This pilot signal is a repeating PN-type sequence known by the receiver unit. Correlation of the received pilot signal with a local replica of the pilot signal enables the receiver unit to estimate the complex impulse response of the channel and adjust demodulator parameters accordingly. For the IS-95 waveform and system parameters it is not necessary or beneficial to report information on the channel conditions measured by the receiver unit back to the transmitter unit.

Given the ever-growing demand for wireless communication, a higher efficiency, higher performance wireless communications system is desirable. One type of higher performance wireless communications system is a Multiple Input/Multiple Output (MIMO) system that employs multiple transmit antennas to transmit over a propagation channel to multiple receive antennas. As in lower performance systems, the propagation channel in a MIMO system is subject to the deleterious effects of multipath, as well as interference from adjacent antennas. Multipath occurs when a transmitted signal arrives at a receiver unit through multiple propagation paths with differing delays. When signals arrive from multiple propagation paths, components of the signals can combine destructively, which is referred to as "fading." In order to improve the efficiency and decrease the complexity of the MIMO system, information as to the characteristics of the propagation channel can be transmitted back to the transmitter unit in order to precondition the signal before transmission.

Preconditioning the signal can be difficult when the characteristics of the propagation channel change rapidly. The channel response can change with time due to the movement of the receiver unit or changes in the environment surrounding the receiver unit. Given a mobile environment, an optimal performance requires that information regarding channel characteristics, such as fading and interference statistics, be determined and transmitted quickly to the transmitter unit before the channel characteristics change significantly. As delay of the measurement and reporting process increases, the utility of the channel response information decreases. A

present need exists for efficient techniques that will provide rapid determination of the channel characteristics.

SUMMARY OF THE INVENTION

The present invention is directed to a method and apparatus for the measuring and reporting of channel state information in a high efficiency, high performance communications system, comprising the steps of: generating a plurality of pilot signals; transmitting the plurality of pilot signals over a propagation channel between a transmitter unit and a plurality of receiver units, wherein the transmitter unit comprises at least one transmit antenna, each of the plurality of receiver units comprises at least one receive antenna, and the propagation channel comprises a plurality of sub-channels between the transmitter unit and the plurality of receiver units; receiving at least one of the plurality of pilot signals at each of the plurality of receiver units; determining a set of transmission characteristics for at least one of the plurality of sub-channels, wherein the step of determining the set of transmission characteristics uses at least one of the plurality of pilot signals received at each of the plurality of receiver units; reporting an information signal from each of the plurality of receiver units to the transmitter unit, wherein the information signal carries the set of transmission characteristics for at least one of the plurality of sub-channels; and optimizing a set of transmission parameters at the transmitter unit, based on the information signal.

In one aspect of the invention, pilot symbols are transmitted on a plurality of disjoint OFDM sub-channel sets. When the pilot symbols are transmitted on disjoint OFDM sub-channels, the characteristics of the propagation channel can be determined through a set of K sub-channels carrying the pilot symbols, wherein K is less than the number of OFDM sub-channels in the system. In addition to transmitting pilot symbols on disjoint sub-channels, the system can transmit a time-domain pilot sequence that can be used to determine characteristics of the propagation

channel. Along with the generation and transmission of pilot symbols, an aspect of the invention is the compression of the amount of information necessary to reconstruct the characteristics of the propagation channel.

According to one aspect of the present invention, there is provided a method of generating pilots in a wireless multiple-input multiple output (MIMO) communication system, comprising: obtaining at least one pilot symbol for each antenna of a plurality of antennas; obtaining an orthogonal sequence for each antenna in the plurality of antennas; and covering the at least one pilot symbol for each antenna with the orthogonal sequence to obtain a sequence of covered pilot symbols for each antenna to obtain at least one covered pilot symbol for each of the plurality of antennas.

According to another aspect of the present invention, there is provided a wireless communication apparatus comprising: a plurality of antennas; means for obtaining at least one pilot symbol for each antenna in the plurality of antennas; means for obtaining an orthogonal sequence for each antenna in the plurality of antennas; and means for covering the pilot symbol for each antenna with the orthogonal sequence to obtain a sequence of covered pilot symbols for the antenna.

According to still another aspect of the present invention, there is provided a wireless communication apparatus comprising: a plurality of antennas; a plurality of modulators coupled to the plurality of antennas; and a transmit data processor, coupled to the plurality of modulators, that provides a plurality of covered pilot symbol for each antenna, wherein each covered pilot symbol is generated by covering a pilot symbol with an orthogonal sequence.

According to yet another aspect of the present invention, there is provided a method of generating pilots in a wireless multiple-input multiple output (MIMO) communication system, comprising: generating at least one pilot symbol for each antenna of a plurality of antennas; and applying an orthogonal sequence to each pilot symbol for each antenna to obtain orthogonal pilot symbols for each antenna.

According to a further aspect of the present invention, there is provided a wireless communication apparatus comprising: a plurality of antennas; means for generating at least one pilot symbol for each antenna of a plurality of antennas; and means for applying an orthogonal sequence to each pilot symbol for each antenna to obtain orthogonal pilot symbols for each antenna.

According to yet a further aspect of the present invention, there is provided a wireless communication apparatus comprising: a plurality of antennas; a plurality of modulators coupled to the plurality of antennas; and a transmit data processor, coupled to the plurality of modulators, that provides a plurality of orthogonal pilot symbol for each antenna, so that each pilot symbol for each antenna is orthogonal to the pilot symbols of the other antennas.

BRIEF DESCRIPTION OF THE DRAWINGS

The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

FIG. 1A is a diagram of a multiple-input multiple-output (MIMO) communications system;

FIG. 1B is a diagram of a OFDM-based MIMO system with feedback of channel state information;

FIG. 1C is a diagram of an exemplary OFDM pilot signal structure that can be used to estimate the channel state information;

FIG. 2 is a diagram that graphically illustrates a specific example of a transmission from a transmit antenna at a transmitter unit;

FIG. 3 is a block diagram of a data processor and a modulator of the communications system shown in FIG. 1A;

FIGS. 4A and 4B are block diagrams of two versions of a channel data processor that can be used for processing one channel data steam such as control, broadcast, voice, or traffic data;

FIGS. 5A through 5C are block diagrams of the processing units that can be used to generate the transmit signal shown in FIG. 2;

FIG. 6 is a block diagram of a receiver unit, having multiple receive antennas, which can be used to receive one or more channel data streams; and

FIG. 7 shows plots that illustrate the spectral efficiency achievable with some of the operating modes of a communications system in accordance with one embodiment.

DETAILED DESCRIPTION OF THE SPECIFIC EMBODIMENTS

FIG. 1A is a diagram of a Multiple Input/Multiple Output (MIMO) communications system 100 capable of implementing some embodiments of the invention. Communications system 100 can be operative to provide a combination of antenna, frequency, and temporal diversity to increase spectral efficiency, improve performance, and enhance flexibility. Increased spectral efficiency is characterized by the ability to transmit more bits per second per Hertz (bps/Hz) when and where possible to better utilize the available system bandwidth. Techniques to obtain higher spectral efficiency are described in further detail below. Improved performance may be quantified, for example, by a lower bit-error-rate (BER) or frame-error-rate (FER) for a given link carrier-to-noise-plus-interference ratio (C/I). And enhanced flexibility is characterized by the ability to accommodate multiple users having different and typically disparate requirements. These goals may be achieved, in part, by employing multi-carrier modulation, time division multiplexing (TDM), multiple transmit and/or receive antennas, and other techniques. The features, aspects, and advantages of the invention are described in further detail below.

As shown in FIG. 1A, communications system 100 includes a first system 110 in communication with a second system 120. System 110 includes a (transmit) data processor 112 that (1) receives or generates data, (2) processes the data to provide antenna, frequency, or temporal diversity, or a combination thereof, and (3) provides processed modulation symbols to a number of modulators (MOD) 114a through 114t. Each modulator 114 further processes the modulation symbols and generates an RF modulated signal suitable for transmission. The RF modulated signals from modulators 114a through 114t are then transmitted from respective antennas 116a through 116t over communications links 118 to system 120.

In FIG. 1A, system 120 includes a number of receive antennas 122a through 122r that receive the transmitted signals and provide the received signals to respective demodulators (DEMOD) 124a through 124r. As shown in FIG. 1A, each receive antenna 122 may receive signals from one or more

transmit antennas 116 depending on a number of factors such as, for example, the operating mode used at system 110, the directivity of the transmit and receive antennas, the characteristics of the communications links, and others. Each demodulator 124 demodulates the respective received signal using a demodulation scheme that is complementary to the modulation scheme used at the transmitter. The demodulated symbols from demodulators 124a through 124r are then provided to a (receive) data processor 126 that further processes the symbols to provide the output data. The data processing at the transmitter and receiver units is described in further detail below.

FIG. 1A shows only the forward link transmission from system 110 to system 120. This configuration may be used for data broadcast and other one-way data transmission applications. In a bi-directional communications system, a reverse link from system 120 to system 110 is also provided, although not shown in FIG. 1A for simplicity. For the bi-directional communications system, each of systems 110 and 120 may operate as a transmitter unit or a receiver unit, or both concurrently, depending on whether data is being transmitted from, or received at, the unit.

For simplicity, communications system 100 is shown to include one transmitter unit (i.e., system 110) and one receiver unit (i.e., system 120). However, in general, multiple transmit antennas and multiple receive antennas are present on each transmitter unit and each receiver unit. The communications system of the invention may include any number of transmitter units and receiver units.

Each transmitter unit may include a single transmit antenna or a number of transmit antennas, such as that shown in FIG. 1A. Similarly, each receiver unit may include a single receive antenna or a number of receive antennas, again such as that shown in FIG. 1A. For example, the communications system may include a central system (i.e., similar to a base station in the IS-95 CDMA system) having a number of antennas that transmit data to, and receive data from, a number of remote systems (i.e., subscriber units, similar to remote stations in the CDMA system), some of

which may include one antenna and others of which may include multiple antennas.

As used herein, an antenna refers to a collection of one or more antenna elements that are distributed in space. The antenna elements may be physically located at a single site or distributed over multiple sites. Antenna elements physically co-located at a single site may be operated as an antenna array (e.g., such as for a CDMA base station). An antenna network consists of a collection of antenna arrays or elements that are physically separated (e.g., several CDMA base stations). An antenna array or an antenna network may be designed with the ability to form beams and to transmit multiple beams from the antenna array or network. For example, a CDMA base station may be designed with the capability to transmit up to three beams to three different sections of a coverage area (or sectors) from the same antenna array. Thus, the three beams may be viewed as three transmissions from three antennas.

The communications system of the invention can be designed to provide a multi-user, multiple access communications scheme capable of supporting subscriber units having different requirements as well as capabilities. The scheme allows the system's total operating bandwidth, W, (e.g., 1.2288 MHz) to be efficiently shared among different types of services that may have highly disparate data rate, delay, and quality of service (QOS) requirements.

Examples of such disparate types of services include voice services and data services. Voice services are typically characterized by a low data rate (e.g., 8 kbps to 32 kbps), short processing delay (e.g., 3 msec to 100 msec overall one-way delay), and sustained use of a communications channel for an extended period of time. The short delay requirements imposed by voice services typically require a small fraction of the system resources to be dedicated to each voice call for the duration of the call. In contrast, data services are characterized by "bursty" traffics in which variable amounts of data are sent at sporadic times. The amount of data can vary significantly from burst-to-burst and from user-to-user. For high efficiency, the

communications system of the invention can be designed with the capability to allocate a portion of the available resources to voice services as required and the remaining resources to data services. A fraction of the available system resources may also be dedicated for certain data services or certain types of data services.

The distribution of data rates achievable by each subscriber unit can vary widely between some minimum and maximum instantaneous values (e.g., from 200 kbps to over 20 Mbps). The achievable data rate for a particular subscriber unit at any given moment may be influenced by a number of factors such as the amount of available transmit power, the quality of the communications link (i.e., the C/I), the coding scheme, and others. The data rate requirement of each subscriber unit may also vary widely between a minimum value (e.g., 8 kbps, for a voice call) all the way up to the maximum supported instantaneous peak rate (e.g., 20 Mbps for bursty data services).

The percentage of voice and data traffic is typically a random variable that changes over time. In accordance with certain aspects of the invention, to efficiently support both types of services concurrently, the communications system of the invention is designed with the capability to dynamic allocate the available resources based on the amount of voice and data traffic. A scheme to dynamically allocate resources is described below. Another scheme to allocate resources is described in the aforementioned U.S. Patent Application Serial No. 08/963,386.

The communications system of the invention provides the above-described features and advantages, and is capable of supporting different types of services having disparate requirements. The features are achieved by employing antenna, frequency, or temporal diversity, or a combination thereof. Antenna, frequency, or temporal diversity can be independently achieved and dynamically selected.

As used herein, antenna diversity refers to the transmission and/or reception of data over more than one antenna, frequency diversity refers to the transmission of data over more than one sub-band, and temporal

diversity refers to the transmission of data over more than one time period. Antenna, frequency, and temporal diversity may include subcategories. For example, transmit diversity refers to the use of more than one transmit antenna in a manner to improve the reliability of the communications link, receive diversity refers to the use of more than one receive antenna in a manner to improve the reliability of the communications link, and spatial diversity refers to the use of multiple transmit and receive antennas to improve the reliability and/or increase the capacity of the communications link. Transmit and receive diversity can also be used in combination to improve the reliability of the communications link without increasing the link capacity. Various combinations of antenna, frequency, and temporal diversity can thus be achieved and are within the scope of the present invention.

Frequency diversity can be provided by use of a multi-carrier modulation scheme such as orthogonal frequency division multiplexing (OFDM), which allows for transmission of data over various sub-bands of the operating bandwidth. Temporal diversity is achieved by transmitting the data over different times, which can be more easily accomplished with the use of time-division multiplexing (TDM). These various aspects of the communications system of the invention are described in further detail below.

In accordance with an aspect of the invention, antenna diversity is achieved by employing a number of (N_T) transmit antennas at the transmitter unit or a number of (N_R) receive antennas at the receiver unit, or multiple antennas at both the transmitter and receiver units. In a terrestrial communications system (e.g., a cellular system, a broadcast system, an MMDS system, and others), an RF modulated signal from a transmitter unit may reach the receiver unit via a number of transmission paths. The characteristics of the transmission paths typically vary over time based on a number of factors. If more than one transmit or receive antenna is used, and if the transmission paths between the transmit and receive antennas are independent (i.e., uncorrelated), which is generally true to at

least an extent, then the likelihood of correctly receiving the transmitted signal increases as the number of antennas increases. Generally, as the number of transmit and receive antennas increases, diversity increases and performance improves.

Antenna diversity is dynamically provided based on the characteristics of the communications link to provide the required performance. For example, a higher degree of antenna diversity can be provided for some types of communication (e.g., signaling), for some types of services (e.g., voice), for some communications link characteristics (e.g., low C/I), or for some other conditions or considerations.

As used herein, antenna diversity includes transmit diversity and receive diversity. For transmit diversity, data is transmitted over multiple transmit antennas. Typically, additional processing is performed on the data transmitted from the transmit antennas to achieve the desired diversity. For example, the data transmitted from different transmit antennas may be delayed or reordered in time, or coded and interleaved across the available transmit antennas. Also, frequency and temporal diversity may be used in conjunction with the different transmit antennas. For receive diversity, modulated signals are received on multiple receive antennas, and diversity is achieved by simply receiving the signals via different transmission paths.

In accordance with another aspect of the invention, frequency diversity can be achieved by employing a multi-carrier modulation scheme. One such scheme that has numerous advantages is OFDM. With OFDM modulation, the overall transmission channel is essentially divided into a number of (L) parallel sub-channels that are used to transmit the same or different data. The overall transmission channel occupies the total operating bandwidth of W, and each of the sub-channels occupies a sub-band having a bandwidth of W/L and centered at a different center frequency. Each sub-channel has a bandwidth that is a portion of the total operating bandwidth. Each of the sub-channels may also be considered an independent data transmission channel that may be associated with a

particular (and possibly unique) processing, coding, and modulation scheme, as described below.

The data may be partitioned and transmitted over any defined set of two or more sub-bands to provide frequency diversity. For example, the transmission to a particular subscriber unit may occur over sub-channel 1 at time slot 1, sub-channel 5 at time slot 2, sub-channel 2 at time slot 3, and so on. As another example, data for a particular subscriber unit may be transmitted over sub-channels 1 and 2 at time slot 1 (e.g., with the same data being transmitted on both sub-channels), sub-channels 4 and 6 at time slot 2, only sub-channel 2 at time slot 3, and so on. Transmission of data over different sub-channels over time can improve the performance of a communications system experiencing frequency selective fading and channel distortion. Other benefits of OFDM modulation are described below.

In accordance with yet another aspect of the invention, temporal diversity is achieved by transmitting data at different times, which can be more easily accomplished using time division multiplexing (TDM). For data services (and possibly for voice services), data transmission occurs over time slots that may be selected to provide immunity to time dependent degradation in the communications link. Temporal diversity may also be achieved through the use of interleaving.

For example, the transmission to a particular subscriber unit may occur over time slots 1 through x, or on a subset of the possible time slots from 1 through x (e.g., time slots 1, 5, 8, and so on). The amount of data transmitted at each time slot may be variable or fixed. Transmission over multiple time slots improves the likelihood of correct data reception due to, for example, impulse noise and interference.

The combination of antenna, frequency, and temporal diversity allows the communications system of the invention to provide robust performance. Antenna, frequency, and/or temporal diversity improves the likelihood of correct reception of at least some of the transmitted data, which may then be used (e.g., through decoding) to correct for some errors

that may have occurred in the other transmissions. The combination of antenna, frequency, and temporal diversity also allows the communications system to concurrently accommodate different types of services having disparate data rate, processing delay, and quality of service requirements.

The communications system of the invention can be designed and operated in a number of different communications modes, with each communications mode employing antenna, frequency, or temporal diversity, or a combination thereof. The communications modes include, for example, a diversity communications mode and a MIMO communications mode. Various combinations of the diversity and MIMO communications modes can also be supported by the communications system. Also, other communications modes can be implemented and are within the scope of the present invention.

The diversity communications mode employs transmit and/or receive diversity, frequency, or temporal diversity, or a combination thereof, and is generally used to improve the reliability of the communications link. In one implementation of the diversity communications mode, the transmitter unit selects a modulation and coding scheme (i.e., configuration) from a finite set of possible configurations, which are known to the receiver units. For example, each overhead and common channel may be associated with a particular configuration that is known to all receiver units. When using the diversity communications mode for a specific user (e.g., for a voice call or a data transmission), the mode and/or configuration may be known *a priori* (e.g., from a previous set up) or negotiated (e.g., via a common channel) by the receiver unit.

In the diversity communications mode, data is transmitted on one or more sub-channels, from one or more antennas, and at one or more time periods. The allocated sub-channels may be associated with the same antenna, or may be sub-channels associated with different antennas. In a common application of the diversity communications mode, which is also referred to as a "pure" diversity communications mode, data is transmitted from all available transmit antennas to the destination receiver unit. The

pure diversity communications mode can be used in instances where the data rate requirements are low or when the C/I is low, or when both are true.

The MIMO communications mode employs antenna diversity at both ends of the communication link and is generally used to improve both the reliability and increase the capacity of the communications link. The MIMO communications mode may further employ frequency and/or temporal diversity in combination with the antenna diversity. The MIMO communications mode, which may also be referred to herein as the spatial communications mode, employs one or more processing modes to be described below.

The diversity communications mode generally has lower spectral efficiency than the MIMO communications mode, especially at high C/I levels. However, at low to moderate C/I values, the diversity communications mode achieves comparable efficiency and can be simpler to implement. In general, the use of the MIMO communications mode provides greater spectral efficiency when used, particularly at moderate to high C/I values. The MIMO communications mode may thus be advantageously used when the data rate requirements are moderate to high.

The communications system can be designed to concurrently support both diversity and MIMO communications modes. The communications modes can be applied in various manners and, for increased flexibility, may be applied independently on a sub-channel basis. The MIMO communications mode is typically applied to specific users. However, each communications mode may be applied on each sub-channel independently, across a subset of sub-channels, across all sub-channels, or on some other basis. For example, the use of the MIMO communications mode may be applied to a specific user (e.g., a data user) and, concurrently, the use of the diversity communications mode may be applied to another specific user (e.g., a voice user) on a different sub-channel. The diversity communications mode may also be applied, for example, on sub-channels experiencing higher path loss.

The communications system of the invention can also be designed to support a number of processing modes. When the transmitter unit is provided with information indicative of the conditions (i.e., the "state") of the communications links, additional processing can be performed at the transmitter unit to further improve performance and increase efficiency. Full channel state information (CSI) or partial CSI may be available to the transmitter unit. Full CSI includes sufficient characterization of the propagation path (i.e., amplitude and phase) between all pairs of transmit and receive antennas for each sub-band. Full CSI also includes the C/I per sub-band. The full CSI may be embodied in a set of matrices of complex gain values that are descriptive of the conditions of the transmission paths from the transmit antennas to the receive antennas, as described below. Partial CSI may include, for example, the C/I of the sub-band . With full CSI or partial CSI, the transmitter unit pre-conditions the data prior to transmission to receiver unit.

The transmitter unit can precondition the signals presented to the transmit antennas in a way that is unique to a specific receiver unit (e.g., the pre-conditioning is performed for each sub-band assigned to that receiver unit). As long as the channel does not change appreciably from the time it is measured by the receiver unit and subsequently sent back to the transmitter and used to precondition the transmission, the intended receiver unit can demodulate the transmission. In this implementation, a full-CSI based MIMO communication can only be demodulated by the receiver unit associated with the CSI used to precondition the transmitted signals.

In the partial-CSI or no-CSI processing modes, the transmitter unit can employ a common modulation and coding scheme (e.g., on each data channel transmission), which then can be (in theory) demodulated by all receiver units. In the partial-CSI processing mode, a single receiver unit can specify the C/I, and the modulation employed on all antennas can be selected accordingly (e.g., for reliable transmission) for that receiver unit. Other receiver units can attempt to demodulate the transmission and, if they have adequate C/I, may be able to successfully recover the transmission. A common (e.g., broadcast) channel can use a no-CSI processing mode to reach all users.

As an example, assume that the MIMO communications mode is applied to a channel data stream that is transmitted on one particular sub-

channel from four transmit antennas. The channel data stream is demultiplexed into four data sub-streams, one data sub-stream for each transmit antenna. Each data sub-stream is then modulated using a particular modulation scheme (e.g., M-PSK, M-QAM, or other) selected based on the CSI for that sub-band and for that transmit antenna. Four modulation sub-streams are thus generated for the four data sub-streams, with each modulation sub-streams including a stream of modulation symbols. The four modulation sub-streams are then pre-conditioned using the eigenvector matrix, as expressed below in equation (1), to generate pre-conditioned modulation symbols. The four streams of pre-conditioned modulation symbols are respectively provided to the four combiners of the four transmit antennas. Each combiner combines the received pre-conditioned modulation symbols with the modulation symbols for the other sub-channels to generate a modulation symbol vector stream for the associated transmit antenna.

The full-CSI based processing is typically employed in the MIMO communications mode where parallel data streams are transmitted to a specific user on each of the channel eigenmodes for the each of the allocated sub-channels. Similar processing based on full CSI can be performed where transmission on only a subset of the available eigenmodes is accommodated in each of the allocated sub-channels(e.g., to implement beam steering). Because of the cost associated with the full-CSI processing (e.g., increased complexity at the transmitter and receiver units, increased overhead for the transmission of the CSI from the receiver unit to the transmitter unit, and so on), full-CSI processing can be applied in certain instances in the MIMO communications mode where the additional increase in performance and efficiency is justified.

In instances where full CSI is not available, less descriptive information on the transmission path (or partial CSI) may be available and can be used to pre-condition the data prior to transmission. For example, the C/I of each of the sub-channels may be available. The C/I information can then be used to control the transmission from various transmit

antennas to provide the required performance in the sub-channels of interest and increase system capacity.

As used herein, full-CSI based processing modes denote processing modes that use full CSI, and partial-CSI based processing modes denote processing modes that use partial CSI. The full-CSI based processing modes include, for example, the full-CSI MIMO mode that utilizes full-CSI based processing in the MIMO communications mode. The partial-CSI based modes include, for example, the partial-CSI MIMO mode that utilizes partial-CSI based processing in the MIMO communications mode.

In instances where full-CSI or partial-CSI processing is employed to allow the transmitter unit to pre-condition the data using the available channel state information (e.g., the eigenmodes or C/I), feedback information from the receiver unit is required, which uses a portion of the reverse link capacity. Therefore, there is a cost associated with the full-CSI and the partial-CSI based processing modes. The cost should be factored into the choice of which processing mode to employ. The partial-CSI based processing mode requires less overhead and may be more efficient in some instances. The no-CSI based processing mode requires no overhead and may also be more efficient than the full-CSI based processing mode or the partial-CSI based processing mode under some other circumstances.

FIG. 2 is a diagram that graphically illustrates at least some of the aspects of the communications system of the invention. FIG. 2 shows a specific example of a transmission from one of N_T transmit antennas at a transmitter unit. In FIG. 2, the horizontal axis is time and the vertical axis is frequency. In this example, the transmission channel includes 16 sub-channels and is used to transmit a sequence of OFDM symbols, with each OFDM symbol covering all 16 sub-channels (one OFDM symbol is indicated at the top of FIG. 2 and includes all 16 sub-bands). A TDM structure is also illustrated in which the data transmission is partitioned into time slots, with each time slot having the duration of, for example, the length of one modulation symbol (i.e., each modulation symbol is used as the TDM interval).

The available sub-channels can be used to transmit signaling, voice, traffic data, and others. In the example shown in FIG. 2, the modulation symbol at time slot 1 corresponds to pilot data, which is periodically transmitted to assist the receiver units to synchronize and perform channel estimation. Other techniques for distributing pilot data over time and frequency can also be used and are within the scope of the present invention. In addition, it may be advantageous to utilize a particular modulation scheme during the pilot interval if all sub-channels are employed (e.g., a PN code with a chip duration of approximately 1/W). Transmission of the pilot modulation symbol typically occurs at a particular frame rate, which is usually selected to be fast enough to permit accurate tracking of variations in the communications link.

The time slots not used for pilot transmissions can then be used to transmit various types of data. For example, sub-channels 1 and 2 may be reserved for the transmission of control and broadcast data to the receiver units. The data on these sub-channels is generally intended to be received by all receiver units. However, some of the messages on the control channel may be user specific, and can be encoded accordingly.

Voice data and traffic data can be transmitted in the remaining sub-channels. For the example shown in FIG. 2, sub-channel 3 at time slots 2 through 9 is used for voice call 1, sub-channel 4 at time slots 2 through 9 is used for voice call 2, sub-channel 5 at time slots 5 through 9 is used for voice call 3, and sub-channel 6 at time slots 7 through 9 is used for voice call 5.

The remaining available sub-channels and time slots may be used for transmissions of traffic data. In the example shown in FIG. 2, data 1 transmission uses sub-channels 5 through 16 at time slot 2 and sub-channels 7 through 16 at time slot 7, data 2 transmission uses sub-channels 5 through 16 at time slots 3 and 4 and sub-channels 6 through 16 at time slots 5, data 3 transmission uses sub-channels 6 through 16 at time slot 6, data 4 transmission uses sub-channels 7 through 16 at time slot 8, data 5 transmission uses sub-channels 7 through 11 at time slot 9, and data 6 transmission uses sub-channels 12 through 16 at time slot 9. Data 1 through

6 transmissions can represent transmissions of traffic data to one or more receiver units.

The communications system of the invention flexibly supports the transmissions of traffic data. As shown in FIG. 2, a particular data transmission (e.g., data 2) may occur over multiple sub-channels and/or multiple time slots, and multiple data transmissions (e.g., data 5 and 6) may occur at one time slot. A data transmission (e.g., data 1) may also occur over non-contiguous time slots. The system can also be designed to support multiple data transmissions on one sub-channel. For example, voice data may be multiplexed with traffic data and transmitted on a single sub-channel.

The multiplexing of the data transmissions can potentially change from OFDM symbol to symbol. Moreover, the communications mode may be different from user to user (e.g., from one voice or data transmission to other). For example, the voice users may use the diversity communications mode, and the data users may use the MIMO communications modes. This concept can be extended to the sub-channel level. For example, a data user may use the MIMO communications mode in sub-channels that have sufficient C/I and the diversity communications mode in remaining sub-channels.

Antenna, frequency, and temporal diversity may be respectively achieved by transmitting data from multiple antennas, on multiple sub-channels in different sub-bands, and over multiple time slots. For example, antenna diversity for a particular transmission (e.g., voice call 1) may be achieved by transmitting the (voice) data on a particular sub-channel (e.g., sub-channel 1) over two or more antennas. Frequency diversity for a particular transmission (e.g., voice call 1) may be achieved by transmitting the data on two or more sub-channels in different sub-bands (e.g., sub-channels 1 and 2). A combination of antenna and frequency diversity may be obtained by transmitting data from two or more antennas and on two or more sub-channels. Temporal diversity may be achieved by transmitting data over multiple time slots. For example, as shown in FIG. 2, data 1

transmission at time slot 7 is a portion (e.g., new or repeated) of the data 1 transmission at time slot 2.

The same or different data may be transmitted from multiple antennas and/or on multiple sub-bands to obtain the desired diversity. For example, the data may be transmitted on: (1) one sub-channel from one antenna, (2) one sub-channel (e.g., sub-channel 1) from multiple antennas, (3) one sub-channel from all N_T antennas, (4) a set of sub-channels (e.g., sub-channels 1 and 2) from one antenna, (5), a set of sub-channels from multiple antennas, (6) a set of sub-channels from all N_T antennas, or (7) a set of channels from a set of antennas (e.g., sub-channel 1 from antennas 1 and 2 at one time slot, sub-channels 1 and 2 from antenna 2 at another time slot, and so on). Thus, any combination of sub-channels and antennas may be used to provide antenna and frequency diversity.

In accordance with certain embodiments of the invention that provide the most flexibility and are capable of achieving high performance and efficiency, each sub-channel at each time slot for each transmit antenna may be viewed as an independent unit of transmission (i.e., a modulation symbol) that can be used to transmit any type of data such as pilot, signaling, broadcast, voice, traffic data, and others, or a combination thereof (e.g., multiplexed voice and traffic data). In such design, a voice call may be dynamically assigned different sub-channels over time.

Flexibility, performance, and efficiency are further achieved by allowing for independence among the modulation symbols, as described below. For example, each modulation symbol may be generated from a modulation scheme (e.g., M-PSK, M-QAM, and others) that results in the best use of the resource at that particular time, frequency, and space.

A number of constraints may be placed to simplify the design and implementation of the transmitter and receiver units. For example, a voice call may be assigned to a particular sub-channel for the duration of the call, or until such time as a sub-channel reassignment is performed. Also, signaling and/or broadcast data may be designated to some fixed sub-channels (e.g., sub-channel 1 for control data and sub-channel 2 for broadcast

data, as shown in FIG. 2) so that the receiver units know a priori which sub-channels to demodulate to receive the data.

Also, each data transmission channel or sub-channel may be restricted to a particular modulation scheme (e.g., M-PSK, M-QAM) for the duration of the transmission or until such time as a new modulation scheme is assigned. For example, in FIG. 2, voice call 1 on sub-channel 3 may use QPSK, voice call 2 on sub-channel 4 may use 16-QAM, data 1 transmission at time slot 2 may use 8-PSK, data 2 transmission at time slots 3 through 5 may use 16-QAM, and so on.

The use of TDM allows for greater flexibility in the transmission of voice data and traffic data, and various assignments of resources can be contemplated. For example, a user can be assigned one sub-channel for each time slot or, equivalently, four sub-channels every fourth time slot, or some other allocations. TDM allows for data to be aggregated and transmitted at designated time slot(s) for improved efficiency.

If voice activity is implemented at the transmitter, then in the intervals where no voice is being transmitted, the transmitter may assign other users to the sub-channel so that the sub-channel efficiency is maximized. In the event that no data is available to transmit during the idle voice periods, the transmitter can decrease (or turn-off) the power transmitted in the sub-channel, reducing the interference levels presented to other users in the system that are using the same sub-channel in another cell in the network. The same feature can be also extended to the overhead, control, data, and other channels.

Allocation of a small portion of the available resources over a continuous time period typically results in lower delays, and may be better suited for delay sensitive services such as voice. Transmission using TDM can provide higher efficiency, at the cost of possible additional delays. The communications system of the invention can allocate resources to satisfy user requirements and achieve high efficiency and performance.

Measuring and Reporting Channel State Information In A MIMO System

Given the complexity of a system using multiple transmit antennas and multiple receive antennas, with the associated dispersive channel effects, the preferred modulation technique is OFDM, which effectively decomposes the channel into a set of non-interfering narrowband channels, or sub-channels. With proper OFDM signal design, a signal transmitted on one subchannel sees "flat fading", i.e., the channel response is effectively constant over the subchannel bandwidth. The channel state information or CSI includes sufficient characterization of the propagation path (i.e., amplitude and phase) between all pairs of transmit and receive antennas for each sub-channel. CSI also includes the information of the relative levels of interference and noise in each sub-channel, that is known as C/I information. The CSI may be embodied in a set of matrices of complex gain values that are descriptive of the conditions of the transmission paths from the transmit antennas to the receive antennas, as described below. With CSI, the transmitter unit pre-conditions the data prior to transmission to receiver unit.

CSI processing is briefly described below. When the CSI is available at the transmitter unit, a simple approach is to decompose the multi-input/multi-output channel into a set of independent channels. Given the channel transfer function at the transmitters, the left eigenvectors may be used to transmit different data streams. The modulation alphabet used with each eigenvector is determined by the available C/I of that mode, given by the eigenvalues. If **H** is the $N_R \times N_T$ matrix that gives the channel response for the N_T transmitter antenna elements and N_R receiver antenna elements at a specific time, and $\underline{\mathbf{x}}$ is the N_T-vector of inputs to the channel, then the received signal can be expressed as:

$$\underline{\mathbf{y}} = \underline{\mathbf{H}}\underline{\mathbf{x}} + \underline{\mathbf{n}},$$

where $\underline{\mathbf{n}}$ is an N_R-vector representing noise plus interference. The eigenvector decomposition of the Hermitian matrix formed by the product of the channel matrix with its conjugate-transpose can be expressed as:

$$\mathbf{H}^{*}\mathbf{H} = E\ddot{E}E^{*} \quad ,$$

where the symbol * denotes conjugate-transpose, E is the eigenvector matrix, and $\ddot{\mathrm{E}}$ is a diagonal matrix of eigenvalues, both of dimension $N_T x N_T$.The transmitter converts a set of N_T modulation symbols $\underline{\mathbf{b}}$ using the eigenvector matrix E. The transmitted modulation symbols from the N_T transmit antennas can thus be expressed as:

$$\underline{\mathbf{x}} = E\underline{\mathbf{b}} \ .$$

For all antennas, the pre-conditioning can thus be achieved by a matrix multiply operation expressed as:

$$\begin{bmatrix} x_1 \\ x_2 \\ \mathrm{M} \\ x_{N_T} \end{bmatrix} = \begin{bmatrix} e_{11}, & e_{12}, & & e_{1N_T} \\ e_{21}, & e_{22}, & & e_{2N_T} \\ & & & \\ e_{N_T1}, & e_{N_T1}, & & e_{N_TN_T} \end{bmatrix} \bullet \begin{bmatrix} b_1 \\ b_2 \\ \mathrm{M} \\ b_{N_T} \end{bmatrix} \qquad \text{Eq (2)}$$

where b_1, b_2, ... and b_{N_T} are respectively the modulation symbols for a particular sub-channel at transmit antennas 1, 2, ... N_T, where each modulation symbol can be generated using, for example, M-PSK, M-QAM, and so on, as described below;

E = is the eigenvector matrix related to the transmission loss from transmit antennas to the receive antennas; and

x_1, x_2, ... x_{N_T} are the pre-conditioned modulation symbols, which can be expressed as:

$$x_1 = b_1 \bullet e_{11} + b_2 \bullet e_{12} + \ \ldots \ + b_{N_T} \bullet e_{1N_T} \quad ,$$

$$x_2 = b_1 \bullet e_{21} + b_2 \bullet e_{22} + \ \ldots \ + b_{N_T} \bullet e_{2N_T} \quad \text{, and}$$

$$x_{N_T} = b_1 \bullet e_{N_T 1} + b_2 \bullet e_{N_T 2} + \ \dots \ + b_{N_T} \bullet e_{N_T N_T} \ .$$

Since **H*H** is Hermitian, the eigenvector matrix is unitary. Thus, if the elements of $\underline{\mathbf{b}}$ have equal power, the elements of $\underline{\mathbf{x}}$ also have equal power. The received signal may then be expressed as:

$$\underline{\mathbf{y}} = \mathbf{HE}\underline{\mathbf{b}} + \underline{\mathbf{n}} \ .$$

The receiver performs a channel-matched-filter operation, followed by multiplication by the right eigenvectors. The result of the channel-matched-filter operation is the vector $\underline{\mathbf{z}}$, which can be expressed as:

$$\underline{\mathbf{z}} = E^* \mathbf{H}^* \mathbf{H} E \underline{\mathbf{b}} + E^* \mathbf{H}^* \underline{\mathbf{n}} = \ddot{\mathbf{E}} \underline{\mathbf{b}} + \underline{\hat{\mathbf{n}}} \ , \qquad \text{Eq.(2)}$$

where the new noise term has covariance that can be expressed as:

$$E(\underline{\hat{n}\hat{n}^*}) = E(E^* \mathbf{H}^* \underline{\mathbf{nn}}^* \mathbf{H} E) = E^* \mathbf{H}^* \mathbf{H} E = \Lambda \ ,$$

i.e., the noise components are independent with variance given by the eigenvalues. The C/I of the i-th component of $\underline{\mathbf{z}}$ is λ_i, the i-th diagonal element of $\ddot{\mathbf{E}}$.

The transmitter unit can thus select a modulation alphabet (i.e., signal constellation) for each of the eigenvectors based on the C/I that is given by the eigenvalue. Provided that the channel conditions do not change appreciably in the interval between the time the CSI is measured at the receiver and reported and used to precondition the transmission at the transmitter, the performance of the communications system will be equivalent to that of a set of independent AWGN channels with known C/I's.

Such a system is illustrated in FIG. 1B. At step 141, the transmitter unit 140 converts data into multiple data sub-channels. Different QAM constellations are employed, depending upon the SNR of the mode and sub-channel. The data for each sub-channel is preconditioned by the eigenmode

matrix for that sub-channel. At step 142, the preconditioned data for a particular antenna undergoes an inverse-Fast Fourier Transform (IFFT) operation to produce a time-domain signal. At step 143, a cyclic extension or a cyclic prefix is appended to the time-domain signal in order to maintain orthogonality among the OFDM sub-channels in the presence of time-dispersion in the propagation channel. One extended symbol value is generated for each OFDM sub-channel and will be referred to hereafter as an OFDM symbol. At step 144, the OFDM symbols are transmitted from the multiple transmit antennas.

Multiple antennas at a receiver unit 145 receive signals at step 146. At step 147, the received signals undergo a Discrete Fourier Transform (DFT) operation to channelize the received signals. At step 148, the data from each subchannel over all of the receive antennas is processed. At this processing step, information regarding channel characteristics is extracted from the data, and converted into a more compressed format. One compression technique is the use of the conjugate channel response and the eigenmode matrix to reduce the amount of information needed to describe channel characteristics. At step 149, a message containing the compressed channel state information is transmitted from the receiver unit 145 to the transmitter unit 140, which will then be used to precondition further transmissions.

To facilitate the derivation of the CSI, the transmit waveform is made up of known pilot symbols for an initial preamble. The pilot waveforms for different transmit antennas comprise disjoint sets of OFDM subchannels as illustrated for the case when $N_t = 4$ in FIG. 1C.

With OFDM modulation, the propagation channel is divided into L parallel sub-channels. In order to determine the CSI quickly, an initial preamble consisting entirely of known symbols is transmitted. In order to efficiently distinguish the differing channel responses of the different transmit-receive antenna patterns, the pilot signals are assigned disjoint subsets of sub-channels. FIG. 1C is a diagram of an exemplary OFDM pilot structure composed of disjoint sub-channel subsets. A sub-channel set

composed of sub-channels {0, 1, 2, . . ., 2^n-1} is decomposed into four disjoint sub-channel subsets A = {0, 4, 8, . . ., 2^n-4}, B = {1, 5, 9, . . . , 2^n-3}, C = {2, 6, 10, . . ., 2^n-2} and D = {3, 7, 11, . . ., 2^n-1}. Sub-channel subset A 150 is transmitted on transmit antenna Tx1 151, sub-channel subset B 152 is transmitted on transmit antenna Tx2 153, sub-channel subset C 154 is transmitted on transmit antenna Tx3 155, and sub-channel subset D 156 is transmitted on transmit antenna Tx4 157. Generally, each transmit antenna transmits on every N^{th} sub-channel across the channel so that all sub-channels are disjoint between transmit antennas. Known pilot symbols can be transmitted on all sub-channels in a sub-channel subset. The minimum spacing between the sub-channels used by a particular transmit antenna is a function of the channel parameters. If the channel response has a large delay spread, then a close spacing may be necessary. If the number of antennas is large enough that the required spacing may not be achieved for all users with a single OFDM symbol, then a number of consecutive OFDM symbols may be employed, with each antenna assigned a disjoint subset of sub-channels on one or more of the multiple pilot symbols.

From each transmit antenna at a transmitter unit, the receiver unit receives pilot symbols on disjoint sub-channels and makes determinations as to channel characteristics of the disjoint sub-channels. As discussed previously, the receiver unit may have one or more receive antennas. Suppose $\underline{x} = \{x_i, i = 1, \ldots, K\}$ are the pilot symbol values that are to be transmitted on K pilot sub-channels for a single transmit antenna. The receiver unit will receive the values $y_{ij} = h_{ij}x_i + n_{ij}$, wherein h_{ij} is the complex channel response for the i^{th} pilot sub-channel received at the j^{th} receive antenna, and n_{ij} is noise. From this relationship, the receiver unit can determine noisy estimates of the channel response of K sub-channels of a single transmit antenna. These noisy estimates may be used to derive estimates for all sub-channels of the propagation channel through a number of different methods, such as simple interpolation to more complex estimation using *a priori* information on the channel dispersion and noise level. The estimates may be improved by transmitting pilot symbols over

consecutive OFDM symbols and then averaging the estimates for each consecutive OFDM symbol.

Estimates are generated at each receive antenna for each transmit antenna broadcasting pilot symbols. The CSI for the complete propagation channel can be represented by the set of channel response matrices $\{H_i, i = 1, 2, \ldots, 2^n\}$, where matrix H_i is associated with the i^{th} sub-channel, and the elements of each matrix H_i are $\{h_{ijk}, j = 1,\ldots,N_r, k = 1,\ldots,N_t\}$, the complex channel response values for each of the N_t transmit and N_r receive antennas.

The use of disjoint sub-channel subsets can further be applied in a system wherein multiple links, e.g., a propagation channel from a transmitter unit to one or more receiver units, are located in close proximity. In a system where a base station transmits signals according to sectors, the transmission area of a sector can overlap the transmission area of another sector. In an ideal base station, transmit antennas in each sector transmit signals in a direction that is completely disjoint from the directions assigned to the transmit antennas of the other sectors. Unfortunately, overlapping areas exist in most sectored base stations. Using this embodiment of the invention, all transmit antennas of a base station are assigned disjoint subsets of sub-channels to avoid interference between the sectors of that base station. Similarly, neighboring base stations may also be the cause of significant interference, and disjoint sets of sub-channels may be assigned among base stations.

In general, the computation of the channel response can be made for every link that is assigned a disjoint sub-channel subset, in the same manner as the response is computed for the principle link. However, a reduced amount of CSI from these interfering links may be reported to the transmitter unit. For example, information as to the average total interference level of neighboring links can be transmitted and used to determine the supportable data rate of the principle link. If several interfering links dominate the average total interference level, then the interference information of these links may be reported individually to the

system in order to determine a more efficient grouping of sub-channels in each disjoint sub-channel subset.

Other CSI information that can be conveyed to the transmitter unit is the total measured power in sub-channels not assigned to the principal link. The total measured power of sub-channels assigned to neighboring links gives an estimate of the total interference plus noise power. If several OFDM symbols are used as the pilot symbol, then the mean measured channel response and the actual received signal values may be used to make a direct estimate of the total noise in a given sub-channel.

In general, the assignment of sub-channels for a network of base stations should follow a pattern of "frequency-reuse," wherein the same sub-channels are used only when the links are sufficiently separated by distance. If a large number of links are interfering with each other, then the number of OFDM sub-channels may be inadequate to allow the assignment of sub-channels for every pilot OFDM symbol. In this circumstance, transmit antennas may be assigned sub-channels for every P-th pilot symbol, where P is an integer value greater than one (1).

In another embodiment of the invention, the OFDM scheme is designed to create OFDM symbol values that minimize or eliminate interference between transmit antennas that use either identical sub-channels or disjoint sub-channels. An orthogonal code, such as Walsh coding, can be used to transform Q pilot signals into Q orthogonal signals representative of the pilot signals. In the case where a Walsh code is used, the number of pilot signals would be a power of two. The use of orthogonal codes can be used together with the previously discussed disjoint sub-channel subsets in order to reduce interference from neighboring links. For example, in a 4×4 MIMO system with a system bandwidth of approximately 1MHz, assume that 256 OFDM sub-channels are to be used. If the multipath is limited to ten microseconds, the disjoint sub-channels carrying pilot symbols should be spaced approximately 50kHz apart or closer. Each sub-channel is approximately 4kHz wide so that a spacing of twelve sub-channels is 48kHz wide. If the OFDM sub-channels are divided into twelve

sets of twenty sub-channels each, sixteen are left unused. Two consecutive OFDM symbols are used as a pilot signal, and orthogonal coding on these two symbols is employed. Hence, there are twenty-four different orthogonal pilot assignments. These twenty-four orthogonal pilots are assigned to different transmit antennas and links to minimize interference.

In another embodiment of the invention, a large number of periodic OFDM symbols can be used as pilot data. The number of OFDM symbols must be large enough so that accurate measurements of interference levels from a large number of different transmit antennas can be made. These average interference levels would be used to set up system-wide constraints on simultaneous transmissions from various sites, i.e., an adaptive blanking scheme to allow all users nearly equivalent performance.

In an alternate embodiment of the invention, the CSI of a MIMO propagation channel can be determined and transmitted for a MIMO system that does not utilize OFDM symbols as pilot signals. Instead, a Maximal-Length Shift Register sequence (m-sequence) can be used to sound the propagation channel. An m-sequence is the output of a shift register with feedback. M-sequences have desirable autocorrelation properties, including the property that correlation over a full period of the sequence with any non-zero circular shift of the sequence yields the value –1, wherein the sequence values are +/-1. Hence, the correlation at zero shift is R, wherein R is the length of the sequence. In order to maintain desirable properties such as correlation in the presence of multipath, a portion of the sequence equal to the delay spread of the channel must be repeated.

For example, if it is known that the channel multipath is limited to some time τ_m and the length of the pilot sequence is at least $R\tau_m$, then R different shifts of the same m-sequence may be used with only minimal mutual interference. These R different shifts are assigned to different transmit antennas of a base station and other base stations that are the cause of major interference.

Links in the MIMO system that are distantly separated can be assigned different m-sequences. The cross-correlation properties of different m-

sequences do not exhibit the minimal correlation properties of a single sequence and its shifts, but different m-sequences behave more or less like random sequences and provide an average correlation level of $\sqrt{R}$ where R is the sequence length. This average correlation level is generally adequate for use in a MIMO system, because of the separation between the links.

A shift register with feedback generates all possible m-sequences, so that sequences are merely shifted versions of a single code word of length $R = 2^m-1$, where m is a positive integer value. Hence, a limited number of different binary m-sequences exist. In order to avoid reuse of the same m-sequence in an area where significant interference may result, filtered versions of longer m-sequences can be used. A filtered version of an m-sequence is no longer binary, but will still display the same basic correlation properties.

For example, suppose that the pilot sequence is to be transmitted at a 1MHz rate, and that the multipath is limited to ten microseconds. Assume that a base station has three sectors, wherein four transmit antennas are assigned to each sector for a total of twelve transmit antennas per site. If a length 127 m-sequence is employed, then twelve different shifts of the sequence may be assigned to the antennas of a single base station, with relative shifts of ten samples each. The total length of the transmitted pilot is then 137 microseconds, which is a full period of the sequence plus ten additional samples to accommodate the multipath spread. Then different base stations can be assigned different m-sequences, with m-sequences repeated in a code reuse pattern designed to minimize the effects of interference from the same m-sequence.

The embodiments of the invention discussed herein have been directed to the design and transmission of pilot signals that will allow a person skilled in the art to derive characteristics of the propagation channel and to report such characteristics to the transmission site. However, the full CSI is a large amount of information and also highly redundant. Many methods are available for compressing the amount of CSI information to be transmitted. One method discussed previously is the use of the Hermatian

matrix H*H, wherein H is the channel response as determined at the receiver unit. The Hermation matrix H*H can be reported to the transmitter unit and be used to precondition transmissions. Due to the properties of Hermitian matrices, only half of the matrix elements need to be transmitted, such as the complex lower triangular portion of the matrix H*H, and the real-valued diagonal. Additional efficiencies are realized if the number of receive antennas is larger than the number of transmit antennas. Another method to reduce the amount of information transmitted to the transmitter unit on the reverse link is to report only a subset of the channel response matrices H_i to the transmitter unit, from which the unreported channel response matrices can be determined through interpolation schemes. In another method, a functional representation of the channel response across the sub-channels may be derived for each transmit/receive antenna pair, e.g., a polynomial function representative of the channel response can be generated. The coefficients of the polynomial function are then transmitted to the transmitter unit.

As an alternative to these methods for compressing CSI information, one embodiment of the invention is directed to the transmission of a time-domain representation of the channel response, which is the channel impulse response. If a time-domain representation of the channel response is simple, as in cases where there are only two or three multipath components, an inverse FFT can be performed upon the set of channel frequency responses. The inverse FFT operation can be performed for each link between a transmit/receive antenna pair. The resulting channel impulse responses are then translated into a set of amplitudes and delays that are reported to the transmitter.

As discussed previously, there is a cost associated with the transmission of CSI in the reverse link, which is reduced when the above embodiments of the invention are implemented in the MIMO system. Another method for reducing the cost is to select users according to the short term average of their CSI requirements. The CSI requirements change as the channel fades, so improved efficiency on the reverse link is achieved

if users estimate the quantity of CSI required, and inform the base station at intervals that may be periodic or aperiodic, depending on the rate of change of the propagation channel observed by the user. The base station may then include this factor in scheduling the use of the forward and reverse links. Scheduling can be arranged so that users associated with slowly changing propagation channels report less frequently than users associated with quickly changing propagation channels. The base station can also arrange the scheduling to take into account factors such as the number of system users and fairness.

In another aspect of this embodiment of the invention, a time interval can be assigned so that CSI updates in a long transmission period can be adjusted according to the actual changes in the propagation channel. Changes in the propagation channel can be monitored at the receiving site in one of a number of possible ways. For example, the difference between the soft decision on the symbols and the closest QAM constellation value can be determined and used as a criterion, or the relative sizes of decoder metrics can also be used. When the quality of a given criterion falls below a predetermined threshold, an update to the CSI is reported to the transmitter unit.

The overall multipath power-delay profile of a link changes very slowly because the average power observed at various delays remains constant, even though channel fading may occur frequently. Hence, the amount of CSI required to characterize a link can vary substantially from link to link. To optimize performance, the coding of the CSI is tailored to the specific link requirements. If the CSI is sent in frequency-domain form, i.e., a set of channel response matrices which are to be interpolated, then links with little multipath require only a small set of channel response matrices.

Structural Components of a High Efficiency, High Performance Communication System

FIG. 3 is a block diagram of a data processor 112 and modulator 114 of system 110 in FIG. 1A. The aggregate input data stream that includes all data to be transmitted by system 110 is provided to a demultiplexer (DEMUX) 310 within data processor 112. Demultiplexer 310 demultiplexes the input data stream into a number of (K) channel data streams, S_1 through S_k. Each channel data stream may correspond to, for example, a signaling channel, a broadcast channel, a voice call, or a traffic data transmission. Each channel data stream is provided to a respective encoder 312 that encodes the data using a particular encoding scheme.

The encoding may include error correction coding or error detection coding, or both, used to increase the reliability of the link. More specifically, such encoding may include, for example, interleaving, convolutional coding, Turbo coding, Trellis coding, block coding (e.g., Reed-Solomon coding), cyclic redundancy check (CRC) coding, and others. Turbo encoding is described in further detail in U.S. Patent Application Serial No. 09/205,511, filed December 4, 1998 entitled "Turbo Code Interleaver Using Linear Congruential Sequences" and in a document entitled "The cdma2000 ITU-R RTT Candidate Submission," hereinafter referred to as the IS-2000 standard, both of which are incorporated herein by reference.

The encoding can be performed on a per channel basis, i.e., on each channel data stream, as shown in FIG. 3. However, the encoding may also be performed on the aggregate input data stream, on a number of channel data streams, on a portion of a channel data stream, across a set of antennas, across a set of sub-channels, across a set of sub-channels and antennas, across each sub-channel, on each modulation symbol, or on some other unit of time, space, and frequency. The encoded data from encoders 312a through 312k is then provided to a data processor 320 that processes the data to generate modulation symbols.

In one implementation, data processor 320 assigns each channel data stream to one or more sub-channels, at one or more time slots, and on one or more antennas. For example, for a channel data stream corresponding to

a voice call, data processor 320 may assign one sub-channel on one antenna (if transmit diversity is not used) or multiple antennas (if transmit diversity is used) for as many time slots as needed for that call. For a channel data stream corresponding to a signaling or broadcast channel, data processor 320 may assign the designated sub-channel(s) on one or more antennas, again depending on whether transmit diversity is used. Data processor 320 then assigns the remaining available resources for channel data streams corresponding to data transmissions. Because of the bursty nature of data transmissions and the greater tolerance to delays, data processor 320 can assign the available resources such that the system goals of high performance and high efficiency are achieved. The data transmissions are thus "scheduled" to achieve the system goals.

After assigning each channel data stream to its respective time slot(s), sub-channel(s), and antenna(s), the data in the channel data stream is modulated using multi-carrier modulation. OFDM modulation is used to provide numerous advantages. In one implementation of OFDM modulation, the data in each channel data stream is grouped to blocks, with each block having a particular number of data bits. The data bits in each block are then assigned to one or more sub-channels associated with that channel data stream.

The bits in each block are then demultiplexed into separate sub-channels, with each of the sub-channels conveying a potentially different number of bits (i.e., based on C/I of the sub-channel and whether MIMO processing is employed). For each of these sub-channels, the bits are grouped into modulation symbols using a particular modulation scheme (e.g., M-PSK or M-QAM) associated with that sub-channel. For example, with 16-QAM, the signal constellation is composed of 16 points in a complex plane (i.e., a + j*b), with each point in the complex plane conveying 4 bits of information. In the MIMO processing mode, each modulation symbol in the sub-channel represents a linear combination of modulation symbols, each of which may be selected from a different constellation.

The collection of L modulation symbols form a modulation symbol vector V of dimensionality L. Each element of the modulation symbol vector V is associated with a specific sub-channel having a unique frequency or tone on which the modulation symbols is conveyed. The collection of these L modulation symbols are all orthogonal to one another. At each time slot and for each antenna, the L modulation symbols corresponding to the L sub-channels are combined into an OFDM symbol using an inverse fast Fourier transform (IFFT). Each OFDM symbol includes data from the channel data streams assigned to the L sub-channels.

OFDM modulation is described in further detail in a paper entitled "Multicarrier Modulation for Data Transmission : An Idea Whose Time Has Come," by John A.C. Bingham, IEEE Communications Magazine, May 1990, which is incorporated herein by reference.

Data processor 320 thus receives and processes the encoded data corresponding to K channel data streams to provide N_T modulation symbol vectors, V_1 through V_{NT}, one modulation symbol vector for each transmit antenna. In some implementations, some of the modulation symbol vectors may have duplicate information on specific sub-channels intended for different transmit antennas. The modulation symbol vectors V_1 through V_{NT} are provided to modulators 114a through 114t, respectively.

In FIG. 3, each modulator 114 includes an IFFT 330, cycle prefix generator 332, and an upconverter 334. IFFT 330 converts the received modulation symbol vectors into their time-domain representations called OFDM symbols. IFFT 330 can be designed to perform the IFFT on any number of sub-channels (e.g., 8, 16, 32, and so on). Alternatively, for each modulation symbol vector converted to an OFDM symbol, cycle prefix generator 332 repeats a portion of the time-domain representation of the OFDM symbol to form the transmission symbol for the specific antenna. The cyclic prefix insures that the transmission symbol retains its orthogonal properties in the presence of multipath delay spread, thereby improving performance against deleterious path effects, as described below. The

implementation of IFFT 330 and cycle prefix generator 332 is known in the art and not described in detail herein.

The time-domain representations from each cycle prefix generator 332 (i.e., the transmission symbols for each antenna) are then processed by upconverter 332, converted into an analog signal, modulated to a RF frequency, and conditioned (e.g., amplified and filtered) to generate an RF modulated signal that is then transmitted from the respective antenna 116.

FIG. 3 also shows a block diagram of a data processor 320. The encoded data for each channel data stream (i.e., the encoded data stream, X) is provided to a respective channel data processor 332. If the channel data stream is to be transmitted over multiple sub-channels and/or multiple antennas (without duplication on at least some of the transmissions), channel data processor 332 demultiplexes the channel data stream into a number of (up to $L \bullet N_T$) data sub-streams. Each data sub-stream corresponds to a transmission on a particular sub-channel at a particular antenna. In typical implementations, the number of data sub-streams is less than $L \bullet N_T$ since some of the sub-channels are used for signaling, voice, and other types of data. The data sub-streams are then processed to generate corresponding sub-streams for each of the assigned sub-channels that are then provided to combiners 334. Combiners 334 combine the modulation symbols designated for each antenna into modulation symbol vectors that are then provided as a modulation symbol vector stream. The N_T modulation symbol vector streams for the N_T antennas are then provided to the subsequent processing blocks (i.e., modulators 114).

In a design that provides the most flexibility, best performance, and highest efficiency, the modulation symbol to be transmitted at each time slot, on each sub-channel, can be individually and independently selected. This feature allows for the best use of the available resource over all three dimensions - time, frequency, and space. The number of data bits transmitted by each modulation symbol may thus differ.

FIG. 4A is a block diagram of a channel data processor 400 that can be used for processing one channel data steam. Channel data processor 400 can

be used to implement one channel data processor 332 in FIG. 3. The transmission of a channel data stream may occur on multiple sub-channels (e.g., as for data 1 in FIG. 2) and may also occur from multiple antennas. The transmission on each sub-channel and from each antenna can represent non-duplicated data.

Within channel data processor 400, a demultiplexer 420 receives and demultiplexes the encoded data stream, X_i, into a number of sub-channel data streams, $X_{i,1}$ through $X_{i,M}$, one sub-channel data stream for each sub-channel being used to transmit data. The data demultiplexing can be uniform or non-uniform. For example, if some information about the transmission paths is known (i.e., full CSI or partial CSI is known), demultiplexer 420 may direct more data bits to the sub-channels capable of transmitting more bps/Hz. However, if no CSI is known, demultiplexer 420 may uniformly directs approximately equal numbers of bits to each of the allocated sub-channels.

Each sub-channel data stream is then provided to a respective spatial division processor 430. Each spatial division processor 430 may further demultiplex the received sub-channel data stream into a number of (up to N_T) data sub-streams, one data sub-stream for each antenna used to transmit the data. Thus, after demultiplexer 420 and spatial division processor 430, the encoded data stream X_i may be demultiplexed into up to $L \bullet N_T$ data sub-streams to be transmitted on up to L sub-channels from up to N_T antennas.

At any particular time slot, up to N_T modulation symbols may be generated by each spatial division processor 430 and provided to N_T combiners 400a through 440t. For example, spatial division processor 430a assigned to sub-channel 1 may provide up to N_T modulation symbols for sub-channel 1 of antennas 1 through N_T. Similarly, spatial division processor 430k assigned to sub-channel k may provide up to N_T symbols for sub-channel k of antennas 1 through N_T. Each combiner 440 receives the modulation symbols for the L sub-channels, combines the symbols for each time slot into a modulation symbol vector, and provides the modulation

symbol vectors as a modulation symbol vector stream, V, to the next processing stage (e.g., modulator 114).

Channel data processor 400 may also be designed to provide the necessary processing to implement the full-CSI or partial-CSI processing modes described above. The CSI processing may be performed based on the available CSI information and on selected channel data streams, sub-channels, antennas, etc. The CSI processing may also be enabled and disabled selectively and dynamically. For example, the CSI processing may be enabled for a particular transmission and disabled for some other transmissions. The CSI processing may be enabled under certain conditions, for example, when the transmission link has adequate C/I.

Channel data processor 400 in FIG. 4A provides a high level of flexibility. However, such flexibility is typically not needed for all channel data streams. For example, the data for a voice call is typically transmitted over one sub-channel for the duration of the call, or until such time as the sub-channel is reassigned. The design of the channel data processor can be greatly simplified for these channel data streams.

FIG. 4B is a block diagram of the processing that can be employed for one channel data steam such as overhead data, signaling, voice, or traffic data. A spatial division processor 450 can be used to implement one channel data processor 332 in FIG. 3 and can be used to support a channel data stream such as, for example, a voice call. A voice call is typically assigned to one sub-channel for multiple time slots (e.g., voice 1 in FIG. 2) and may be transmitted from multiple antennas. The encoded data stream, X_j, is provided to spatial division processor 450 that groups the data into blocks, with each block having a particular number of bits that are used to generate a modulation symbol. The modulation symbols from spatial division processor 450 are then provided to one or more combiners 440 associated with the one or more antennas used to transmit the channel data stream.

A specific implementation of a transmitter unit capable of generating the transmit signal shown in FIG. 2 is now described for a better

understanding of the invention. At time slot 2 in FIG. 2, control data is transmitted on sub-channel 1, broadcast data is transmitted on sub-channel 2, voice calls 1 and 2 are assigned to sub-channels 3 and 4, respectively, and traffic data is transmitted on sub-channels 5 through 16. In this example, the transmitter unit is assumed to include four transmit antennas (i.e., $N_T = 4$) and four transmit signals (i.e., four RF modulated signals) are generated for the four antennas.

FIG. 5A is a block diagram of a portion of the processing units that can be used to generate the transmit signal for time slot 2 in FIG. 2. The input data stream is provided to a demultiplexer (DEMUX) 510 that demultiplexes the stream into five channel data streams, S_1 through S_5, corresponding to control, broadcast, voice 1, voice 2, and data 1 in FIG. 2. Each channel data stream is provided to a respective encoder 512 that encodes the data using an encoding scheme selected for that stream.

In this example, channel data streams S_1 through S_3 are transmitted using transmit diversity. Thus, each of the encoded data streams X_1 through X_3 is provided to a respective channel data processor 532 that generates the modulation symbols for that stream. The modulation symbols from each of the channel data processors 532a through 532c are then provided to all four combiners 540a through 540d. Each combiner 540 receives the modulation symbols for all 16 sub-channels designated for the antenna associated with the combiner, combines the symbols on each sub-channel at each time slot to generate a modulation symbol vector, and provides the modulation symbol vectors as a modulation symbol vector stream, V, to an associated modulator 114. As indicated in FIG. 5A, channel data stream S_1 is transmitted on sub-channel 1 from all four antennas, channel data stream S_2 is transmitted on sub-channel 2 from all four antennas, and channel data stream S_3 is transmitted on sub-channel 3 from all four antennas.

FIG. 5B is a block diagram of a portion of the processing units used to process the encoded data for channel data stream S_4. In this example, channel data stream S_4 is transmitted using spatial diversity (and not transmit diversity as used for channel data streams S_1 through S_3). With

spatial diversity, data is demultiplexed and transmitted (concurrently in each of the assigned sub-channels or over different time slots) over multiple antennas. The encoded data stream X_4 is provided to a channel data processor 532d that generates the modulation symbols for that stream. The modulation symbols in this case are linear combinations of modulation symbols selected from symbol alphabets that correspond to each of the eigenmodes of the channel. In this example, there are four distinct eigenmodes, each of which is capable of conveying a different amount of information. As an example, suppose eigenmode 1 has a C/I that allows 64-QAM (6 bits) to be transmitted reliably, eigenmode 2 permits 16-QAM (4 bits), eigenmode 3 permits QPSK (2 bits) and eigenmode 4 permits BPSK (1 bit) to be used. Thus, the combination of all four eigenmodes allows a total of 13 information bits to be transmitted simultaneously as an effective modulation symbol on all four antennas in the same sub-channel. The effective modulation symbol for the assigned sub-channel on each antenna is a linear combination of the individual symbols associated with each eigenmode, as described by the matrix multiply given in equation (1) above.

FIG. 5C is a block diagram of a portion of the processing units used to process channel data stream S_5. The encoded data stream X_5 is provided to a demultiplexer (DEMUX) 530 that demultiplexes the stream X_5 into twelve sub-channel data streams, $X_{5,11}$ through $X_{5,16}$, one sub-channel data stream for each of the allocated sub-channels 5 through 16. Each sub-channel data stream is then provided to a respective sub-channel data processor 536 that generates the modulation symbols for the associated sub-channel data stream. The sub-channel symbol stream from sub-channel data processors 536a through 536l are then provided to demultiplexers 538a through 538l, respectively. Each demultiplexer 538 demultiplexes the received sub-channel symbol stream into four symbol sub-streams, with each symbol sub-stream corresponding to a particular sub-channel at a particular antenna. The four symbol sub-streams from each demultiplexer 538 are then provided to the four combiners 540a through 540d.

In FIG. 5C, a sub-channel data stream is processed to generate a sub-channel symbol stream that is then demultiplexed into four symbol sub-streams, one symbol sub-stream for a particular sub-channel of each antenna. This implementation is a different from that described for FIG. 4A. In FIG. 4A, the sub-channel data stream designated for a particular sub-channel is demultiplexed into a number of data sub-streams, one data sub-stream for each antenna, and then processed to generate the corresponding symbol sub-streams. The demultiplexing in FIG. 5C is performed after the symbol modulation whereas the demultiplexing in FIG. 4A is performed before the symbol modulation. Other implementations may also be used and are within the scope of the present invention.

Each combination of sub-channel data processor 536 and demultiplexer 538 in FIG. 5C performs in similar manner as the combination of sub-channel data processor 532d and demultiplexer 534d in FIG. 5B. The rate of each symbol sub-stream from each demultiplexer 538 is, on the average, a quarter of the rate of the symbol stream from the associated channel data processor 536.

FIG. 6 is a block diagram of a receiver unit 600, having multiple receive antennas, which can be used to receive one or more channel data streams. One or more transmitted signals from one or more transmit antennas can be received by each of antennas 610a through 610r and routed to a respective front end processor 612. For example, receive antenna 610a may receive a number of transmitted signals from a number of transmit antennas, and receive antenna 610r may similarly receive multiple transmitted signals. Each front end processor 612 conditions (e.g., filters and amplifies) the received signal, downconverts the conditioned signal to an intermediate frequency or baseband, and samples and quantizes the downconverted signal. Each front end processor 612 typically further demodulates the samples associated with the specific antenna with the received pilot to generate "coherent" samples that are then provided to a respective FFT processor 614, one for each receive antenna.

Each FFT processor 614 generates transformed representations of the received samples and provides a respective stream of modulation symbol vectors. The modulation symbol vector streams from FFT processors 614a through 614r are then provided to demultiplexer and combiners 620, which channelizes the stream of modulation symbol vectors from each FFT processor 614 into a number of (up to L) sub-channel symbol streams. The sub-channel symbol streams from all FFT processors 614 are then processed, based on the (e.g., diversity or MIMO) communications mode used, prior to demodulation and decoding.

For a channel data stream transmitted using the diversity communications mode, the sub-channel symbol streams from all antennas used for the transmission of the channel data stream are presented to a combiner that combines the redundant information across time, space, and frequency. The stream of combined modulation symbols are then provided to a (diversity) channel processor 630 and demodulated accordingly.

For a channel data stream transmitted using the MIMO communications mode, all sub-channel symbol streams used for the transmission of the channel data stream are presented to a MIMO processor that orthogonalizes the received modulation symbols in each sub-channel into the distinct eigenmodes. The MIMO processor performs the processing described by equation (2) above and generates a number of independent symbol sub-streams corresponding to the number of eigenmodes used at the transmitter unit. For example, MIMO processor can perform multiplication of the received modulation symbols with the left eigenvectors to generate post-conditioned modulation symbols, which correspond to the modulation symbols prior to the full-CSI processor at the transmitter unit. The (post-conditioned) symbol sub-streams are then provided to a (MIMO) channel processor 630 and demodulated accordingly. Thus, each channel processor 630 receives a stream of modulation symbols (for the diversity communications mode) or a number of symbol sub-streams (for the MIMO communications mode). Each stream or sub-stream of modulation symbols is then provided to a respective demodulator (DEMOD) that implements a

demodulation scheme (e.g., M-PSK, M-QAM, or others) that is complementary to the modulation scheme used at the transmitter unit for the sub-channel being processed. For the MIMO communications mode, the demodulated data from all assigned demodulators may then be decoded independently or multiplexed into one channel data stream and then decoded, depending upon the coding and modulation method employed at the transmitter unit. For both the diversity and MIMO communications modes, the channel data stream from channel processor 630 may then provided to a respective decoder 640 that implements a decoding scheme complementary to that used at the transmitter unit for the channel data stream. The decoded data from each decoder 540 represents an estimate of the transmitted data for that channel data stream.

FIG. 6 represents one embodiment of a receiver unit. Other designs can contemplated and are within the scope of the present invention. For example, a receiver unit may be designed with only one receive antenna, or may be designed capable of simultaneously processing multiple (e.g., voice, data) channel data streams.

As noted above, multi-carrier modulation is used in the communications system of the invention. In particular, OFDM modulation can be employed to provide a number of benefits including improved performance in a multipath environment, reduced implementation complexity (in a relative sense, for the MIMO mode of operation), and flexibility. However, other variants of multi-carrier modulation can also be used and are within the scope of the present invention.

OFDM modulation can improve system performance due to multipath delay spread or differential path delay introduced by the propagation environment between the transmitting antenna and the receiver antenna. The communications link (i.e., the RF channel) has a delay spread that may potentially be greater than the reciprocal of the system operating bandwidth, W. Because of this, a communications system employing a modulation scheme that has a transmit symbol duration of less than the delay spread will experience inter-symbol interference (ISI). The ISI

distorts the received symbol and increases the likelihood of incorrect detection.

With OFDM modulation, the transmission channel (or operating bandwidth) is essentially divided into a (large) number of parallel sub-channels (or sub-bands) that are used to communicate the data. Because each of the sub-channels has a bandwidth that is typically much less than the coherence bandwidth of the communications link, ISI due to delay spread in the link is significantly reduced or eliminated using OFDM modulation. In contrast, most conventional modulation schemes (e.g., QPSK) are sensitive to ISI unless the transmission symbol rate is small compared to the delay spread of the communications link.

As noted above, cyclic prefixes can be used to combat the deleterious effects of multipath. A cyclic prefix is a portion of an OFDM symbol (usually the front portion, after the IFFT) that is wrapped around to the back of the symbol. The cyclic prefix is used to retain orthogonality of the OFDM symbol, which is typically destroyed by multipath.

As an example, consider a communications system in which the channel delay spread is less than 10 μsec. Each OFDM symbol has appended onto it a cyclic prefix that insures that the overall symbol retains its orthogonal properties in the presence of multipath delay spread. Since the cyclic prefix conveys no additional information, it is essentially overhead. To maintain good efficiency, the duration of the cyclic prefix is selected to be a small fraction of the overall transmission symbol duration. For the above example, using a 5% overhead to account for the cyclic prefix, a transmission symbol duration of 200 μsec is adequate for a 10 μsec maximum channel delay spread. The 200 μsec transmission symbol duration corresponds to a bandwidth of 5 kHz for each of the sub-bands. If the overall system bandwidth is 1.2288 MHz, 250 sub-channels of approximately 5 kHz can be provided. In practice, it is convenient for the number of sub-channels to be a power of two. Thus, if the transmission symbol duration is increased to

205 μsec and the system bandwidth is divided into M = 256 sub-bands, each sub-channel will have a bandwidth of 4.88 kHz.

In certain embodiments of the invention, OFDM modulation can reduce the complexity of the system. When the communications system incorporates MIMO technology, the complexity associated with the receiver unit can be significant, particularly when multipath is present. The use of OFDM modulation allows each of the sub-channels to be treated in an independent manner by the MIMO processing employed. Thus, OFDM modulation can significantly simplify the signal processing at the receiver unit when MIMO technology is used.

OFDM modulation can also afford added flexibility in sharing the system bandwidth, W, among multiple users. Specifically, the available transmission space for OFDM symbols can be shared among a group of users. For example, low rate voice users can be allocated a sub-channel or a fraction of a sub-channel in OFDM symbol, while the remaining sub-channels can be allocated to data users based on aggregate demand. In addition, overhead, broadcast, and control data can be conveyed in some of the available sub-channels or (possibly) in a portion of a sub-channel.

As described above, each sub-channel at each time slot is associated with a modulation symbol that is selected from some alphabet such as M-PSK or M-QAM. In certain embodiments, the modulation symbol in each of the L sub-channels can be selected such that the most efficient use is made of that sub-channel. For example, sub-channel 1 can be generated using QPSK, sub-channel 2 can be generate using BPSK, sub-channel 3 can be generated using 16-QAM, and so on. Thus, for each time slot, up to L modulation symbols for the L sub-channels are generated and combined to generate the modulation symbol vector for that time slot.

One or more sub-channels can be allocated to one or more users. For example, each voice user may be allocated a single sub-channel. The remaining sub-channels can be dynamically allocated to data users. In this case, the remaining sub-channels can be allocated to a single data user or divided among multiple data users. In addition, some sub-channels can be

reserved for transmitting overhead, broadcast, and control data. In certain embodiments of the invention, it may be desirable to change the sub-channel assignment from (possibly) modulation symbol to symbol in a pseudo-random manner to increase diversity and provide some interference averaging.

In a CDMA system, the transmit power on each reverse link transmission is controlled such that the required frame error rate (FER) is achieved at the base station at the minimal transmit power, thereby minimizing interference to other users in the system. On the forward link of the CDMA system, the transmit power is also adjusted to increase system capacity.

In the communications system of the invention, the transmit power on the forward and reverse links can be controlled to minimize interference and maximize system capacity. Power control can be achieved in various manners. For example, power control can be performed on each channel data stream, on each sub-channel, on each antenna, or on some other unit of measurement. When operating in the diversity communications mode, if the path loss from a particular antenna is great, transmission from this antenna can be reduced or muted since little may be gained at the receiver unit. Similarly, if transmission occurs over multiple sub-channels, less power may be transmitted on the sub-channel(s) experiencing the most path loss.

In an implementation, power control can be achieved with a feedback mechanism similar to that used in the CDMA system. Power control information can be sent periodically or autonomously from the receiver unit to the transmitter unit to direct the transmitter unit to increase or decrease its transmit power. The power control bits may be generated based on, for example, the BER or FER at the receiver unit.

FIG. 7 shows plots that illustrate the spectral efficiency associated with some of the communications modes of the communications system of the invention. In FIG. 7, the number of bits per modulation symbol for a given bit error rate is given as a function of C/I for a number of system

configurations. The notation $N_T x N_R$ denotes the dimensionality of the configuration, with N_T = number of transmit antennas and N_R = number of receive antennas. Two diversity configurations, namely 1x2 and 1x4, and four MIMO configurations, namely 2x2, 2x4, 4x4, and 8x4, are simulated and the results are provided in FIG. 7.

As shown in the plots, the number of bits per symbol for a given BER ranges from less than 1 bps/Hz to almost 20 bps/Hz. At low values of C/I, the spectral efficiency of the diversity communications mode and MIMO communications mode is similar, and the improvement in efficiency is less noticeable. However, at higher values of C/I, the increase in spectral efficiency with the use of the MIMO communications mode becomes more dramatic. In certain MIMO configurations and for certain conditions, the instantaneous improvement can reach up to 20 times.

From these plots, it can be observed that spectral efficiency generally increases as the number of transmit and receive antennas increases. The improvement is also generally limited to the lower of N_T and N_R. For example, the diversity configurations, 1x2 and 1x4, both asymptotically reach approximately 6 bps/Hz.

In examining the various data rates achievable, the spectral efficiency values given in FIG. 7 can be applied to the results on a sub-channel basis to obtain the range of data rates possible for the sub-channel. As an example, for a subscriber unit operating at a C/I of 5 dB, the spectral efficiency achievable for this subscriber unit is between 1 bps/Hz and 2.25 bps/Hz, depending on the communications mode employed. Thus, in a 5 kHz sub-channel, this subscriber unit can sustain a peak data rate in the range of 5 kbps to 10.5 kbps. If the C/I is 10 dB, the same subscriber unit can sustain peak data rates in the range of 10.5 kbps to 25 kbps per sub-channel. With 256 sub-channels available, the peak sustained data rate for a subscriber unit operating at 10 dB C/I is then 6.4 Mbps. Thus, given the data rate requirements of the subscriber unit and the operating C/I for the subscriber unit, the system can allocate the necessary number of sub-channels to meet the requirements. In the case of data services, the number of sub-channels

allocated per time slot may vary depending on, for example, other traffic loading.

The reverse link of the communications system can be designed similar in structure to the forward link. However, instead of broadcast and common control channels, there may be random access channels defined in specific sub-channels or in specific modulation symbol positions of the frame, or both. These may be used by some or all subscriber units to send short requests (e.g., registration, request for resources, and so on) to the central station. In the common access channels, the subscriber units may employ common modulation and coding. The remaining channels may be allocated to separate users as in the forward link. Allocation and de-allocation of resources (on both the forward and reverse links) can be controlled by the system and can be communicated on the control channel in the forward link.

One design consideration for on the reverse link is the maximum differential propagation delay between the closest subscriber unit and the furthest subscriber unit. In systems where this delay is small relative to the cyclic prefix duration, it may not be necessary to perform correction at the transmitter unit. However, in systems in which the delay is significant, the cyclic prefix can be extended to account for the incremental delay. In some instances, it may be possible to make a reasonable estimate of the round trip delay and correct the time of transmit so that the symbol arrives at the central station at the correct instant. Usually there is some residual error, so the cyclic prefix may also further be extended to accommodate this residual error.

In the communications system, some subscriber units in the coverage area may be able to receive signals from more than one central station. If the information transmitted by multiple central stations is redundant on two or more sub-channels and/or from two or more antennas, the received signals can be combined and demodulated by the subscriber unit using a diversity-combining scheme. If the cyclic prefix employed is sufficient to handle the differential propagation delay between the earliest and latest

arrival, the signals can be (optimally) combined in the receiver and demodulated correctly. This diversity reception is well known in broadcast applications of OFDM. When the sub-channels are allocated to specific subscriber units, it is possible for the same information on a specific sub-channel to be transmitted from a number of central stations to a specific subscriber unit. This concept is similar to the soft handoff used in CDMA systems.

As shown above, the transmitter unit and receiver unit are each implemented with various processing units that include various types of data processor, encoders, IFFTs, FFTs, demultiplexers, combiners, and so on. These processing units can be implemented in various manners such as an application specific integrated circuit (ASIC), a digital signal processor, a microcontroller, a microprocessor, or other electronic circuits designed to perform the functions described herein. Also, the processing units can be implemented with a general-purpose processor or a specially designed processor operated to execute instruction codes that achieve the functions described herein. Thus, the processing units described herein can be implemented using hardware, software, or a combination thereof.

The foregoing description of the preferred embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without the use of the inventive faculty. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

WHAT IS CLAIMED IS:

CLAIMS:

1. A method of generating pilots in a wireless multiple-input multiple output (MIMO) communication system, comprising:

obtaining at least one pilot symbol for each antenna of a plurality of antennas;

obtaining an orthogonal sequence for each antenna in the plurality of antennas; and

covering the at least one pilot symbol for each antenna with the orthogonal sequence to obtain a sequence of covered pilot symbols for each antenna to obtain at least one covered pilot symbol for each of the plurality of antennas.

2. The method of claim 1, wherein obtaining the orthogonal sequence comprises obtaining a different orthogonal sequence for each antenna.

3. The method of claim 1, further comprising transmitting the plurality of orthogonal pilots from the plurality of antennas.

4. The method of claim 1, wherein the orthogonal sequences are Walsh sequences.

5. The method of claim 4, wherein the Walsh sequences have a length of 1/W chips.

6. The method of claim 1, wherein obtaining at least one pilot symbol comprises obtaining a same at least one pilot symbol for each of the plurality of antennas.

7. The method of claim 1, further comprising transmitting at least one covered pilot symbol on each antenna from a predetermined set of subbands.

8. The method of claim 1, further comprising transmitting the sequence of covered pilot symbols during predetermined time slots.

9. The method of claim 1, further comprising applying a cyclic prefix to each covered pilot symbol.

10. A wireless communication apparatus comprising:

a plurality of antennas;

means for obtaining at least one pilot symbol for each antenna in the plurality of antennas;

means for obtaining an orthogonal sequence for each antenna in the plurality of antennas; and

means for covering the pilot symbol for each antenna with the orthogonal sequence to obtain a sequence of covered pilot symbols for the antenna.

11. The apparatus of claim 10, wherein the means for obtaining the orthogonal sequence comprises means for obtaining a different orthogonal sequence for each antenna.

12. The apparatus of claim 10, wherein the orthogonal sequences are Walsh sequences.

13. The apparatus of claim 12, wherein the Walsh sequences have a length of 1/W chips.

14. The apparatus of claim 10, further comprising means for designating transmission of the covered pilot symbols on a predetermined set of subbands.

15. The apparatus of claim 10, further comprising means for transmitting the sequence of covered pilot symbols during predetermined time slots.

16. The apparatus of claim 10, further comprising means for applying a cyclic prefix to each pilot symbol.

17. A wireless communication apparatus comprising:

a plurality of antennas;

a plurality of modulators coupled to the plurality of antennas; and

a transmit data processor, coupled to the plurality of modulators, that provides a plurality of covered pilot symbol for each antenna, wherein each covered pilot symbol is generated by covering a pilot symbol with an orthogonal sequence.

18. The apparatus of claim 17, wherein a different orthogonal sequence is used for each antenna.

19. The apparatus of claim 17, wherein the orthogonal sequences are Walsh sequences.

20. The apparatus of claim 19, wherein the Walsh sequences have a length of 1/W chips.

21. The apparatus of claim 17, wherein the pilot symbols are transmitted on a predetermined set of subbands.

22. The apparatus of claim 17, wherein the modulators apply a cyclic prefix to each pilot symbol.

23. The apparatus of claim 17, wherein the sequence of covered pilot symbols are transmitted during predetermined time slots.

24. A method of generating pilots in a wireless multiple-input multiple output (MIMO) communication system, comprising:

generating at least one pilot symbol for each antenna of a plurality of antennas; and

applying an orthogonal sequence to each pilot symbol for each antenna to obtain orthogonal pilot symbols for each antenna.

25. The method of claim 24, further comprising generating a different orthogonal sequence for each antenna so that the pilot symbols of each antenna are orthogonal to pilot symbols of each other antenna.

26. The method of claim 24, further comprising transmitting the plurality of orthogonal pilot symbols from the plurality of antennas.

27. The method of claim 24, wherein the orthogonal sequences are Walsh sequences.

28. The method of claim 27, wherein the Walsh sequences have a length of 1/W chips.

29. The method of claim 24, wherein obtaining comprises obtaining a same pilot symbol for the plurality of antennas.

30. The method of claim 24, wherein the pilot symbols are transmitted during predetermined time slots.

31. The method of claim 24, further comprising applying a cyclic prefix to each pilot symbol.

32. A wireless communication apparatus comprising:

a plurality of antennas;

means for generating at least one pilot symbol for each antenna of a plurality of antennas; and

means for applying an orthogonal sequence to each pilot symbol for each antenna to obtain orthogonal pilot symbols for each antenna.

33. The apparatus of claim 32, further comprising means for obtaining a different orthogonal sequence for each antenna.

34. The apparatus of claim 32, wherein the orthogonal sequences are Walsh sequences.

35. The apparatus of claim 34, wherein the Walsh sequences have a length of 1/W chips.

36. The apparatus of claim 32, further comprising means for designating transmission of the orthogonal pilot symbols on a predetermined set of subbands.

37. The apparatus of claim 32, further comprising means for applying a cyclic prefix to each pilot symbol.

38. The apparatus of claim 32, further comprising means for transmitting the orthogonal pilot symbols during predetermined time slots.

39. A wireless communication apparatus comprising:

a plurality of antennas;

a plurality of modulators coupled to the plurality of antennas; and

a transmit data processor, coupled to the plurality of modulators, that provides a plurality of orthogonal pilot symbol for each antenna, so that each pilot symbol for each antenna is orthogonal to the pilot symbols of the other antennas.

40. The apparatus of claim 39, wherein a different orthogonal sequence is used for each antenna.

41. The apparatus of claim 39, wherein the orthogonal sequences are Walsh sequences.

42. The apparatus of claim 41, wherein the Walsh sequences have a length of 1/W chips.

43. The apparatus of claim 39, wherein the pilot symbols are transmitted on a predetermined set of subbands.

44. The apparatus of claim 39, wherein the modulators apply a cyclic prefix to each pilot symbol.

SMART & BIGGAR
OTTAWA, CANADA

PATENT AGENTS

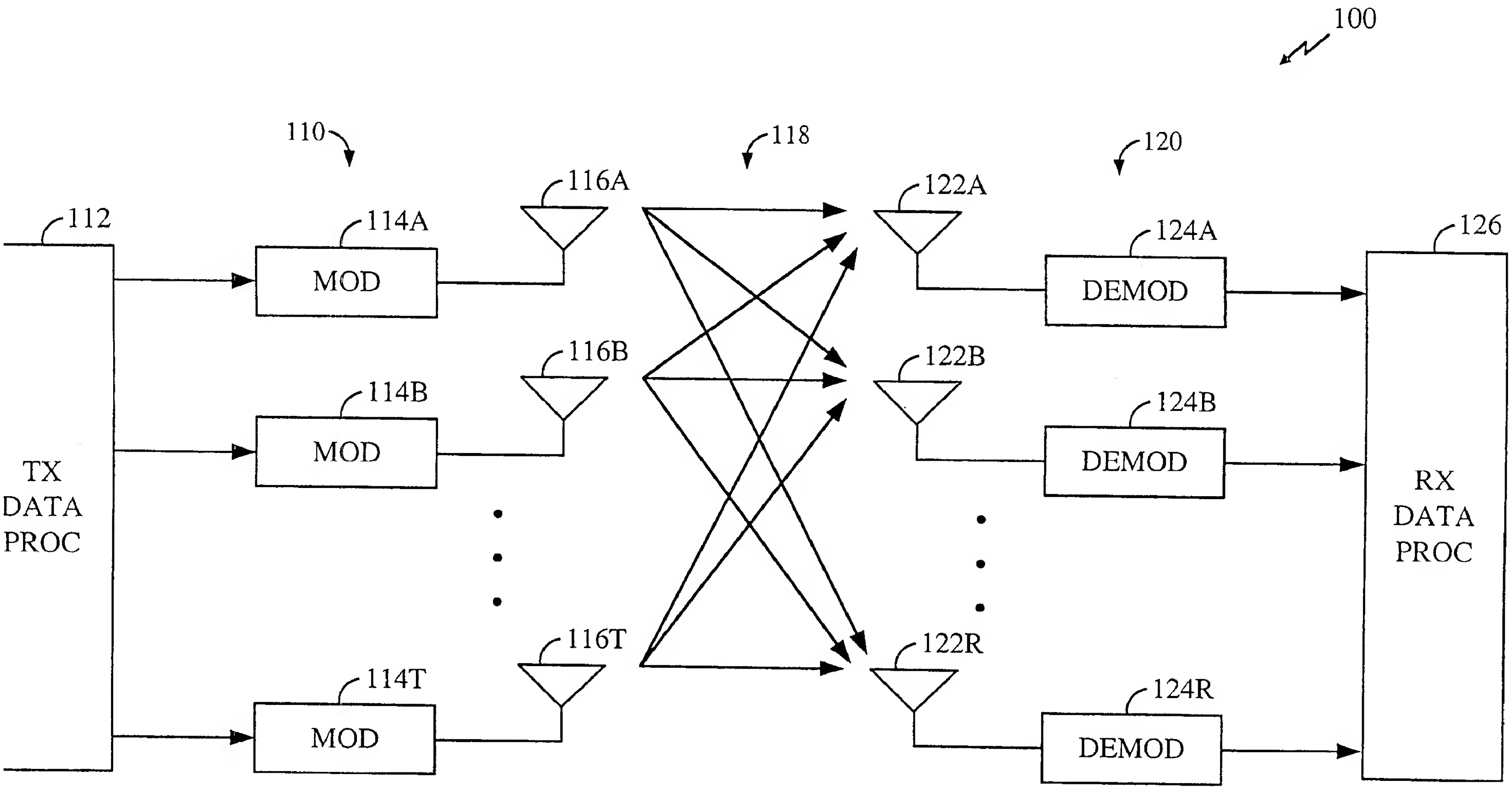

100
110
118
120
112
TX
DATA
PROC
114A
MOD
114B
MOD
114T
MOD
116A
116B
116T
122A
122B
122R
124A
DEMOD
124B
DEMOD
124R
DEMOD
126
RX
DATA
PROC

FIG. 1A

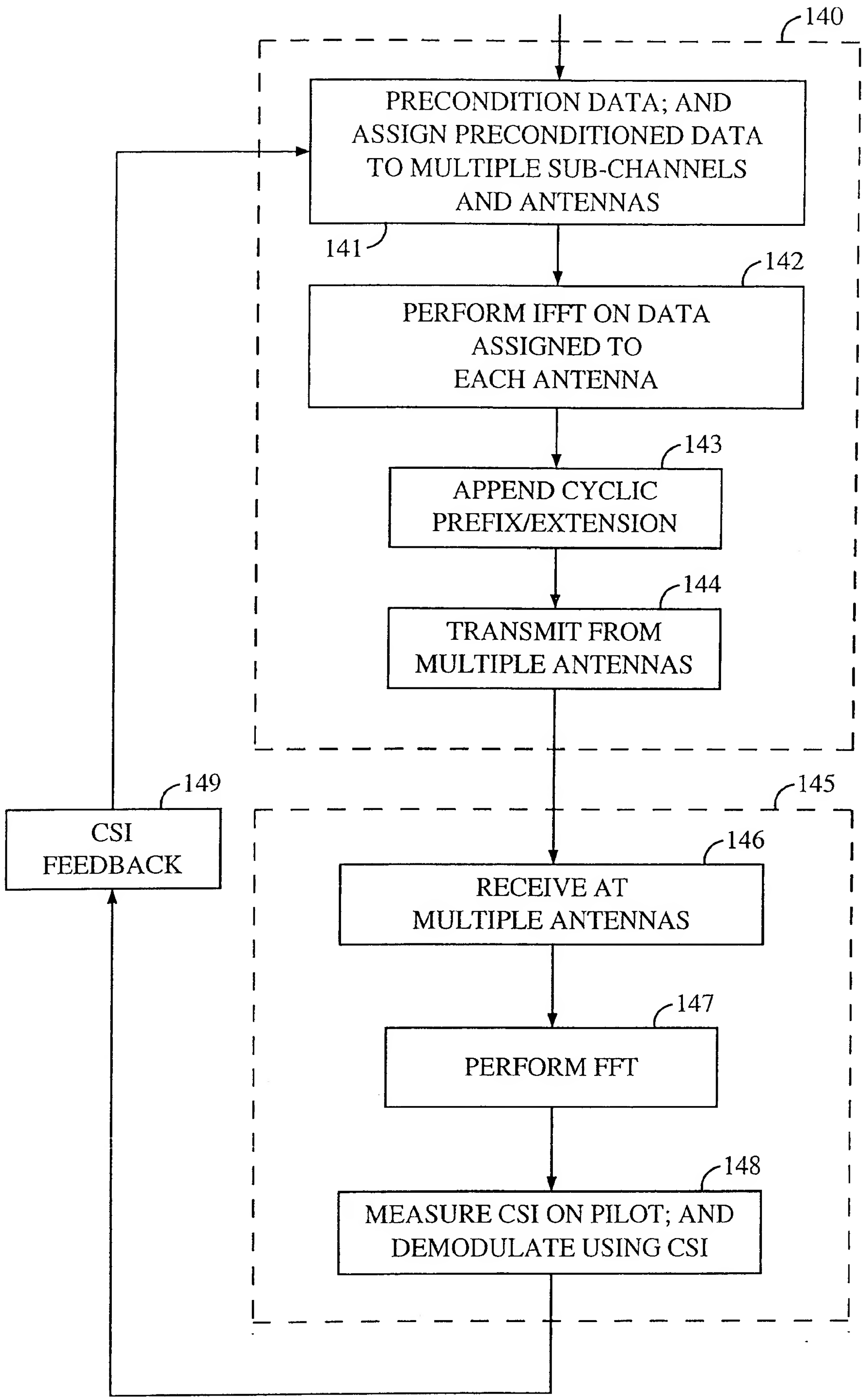
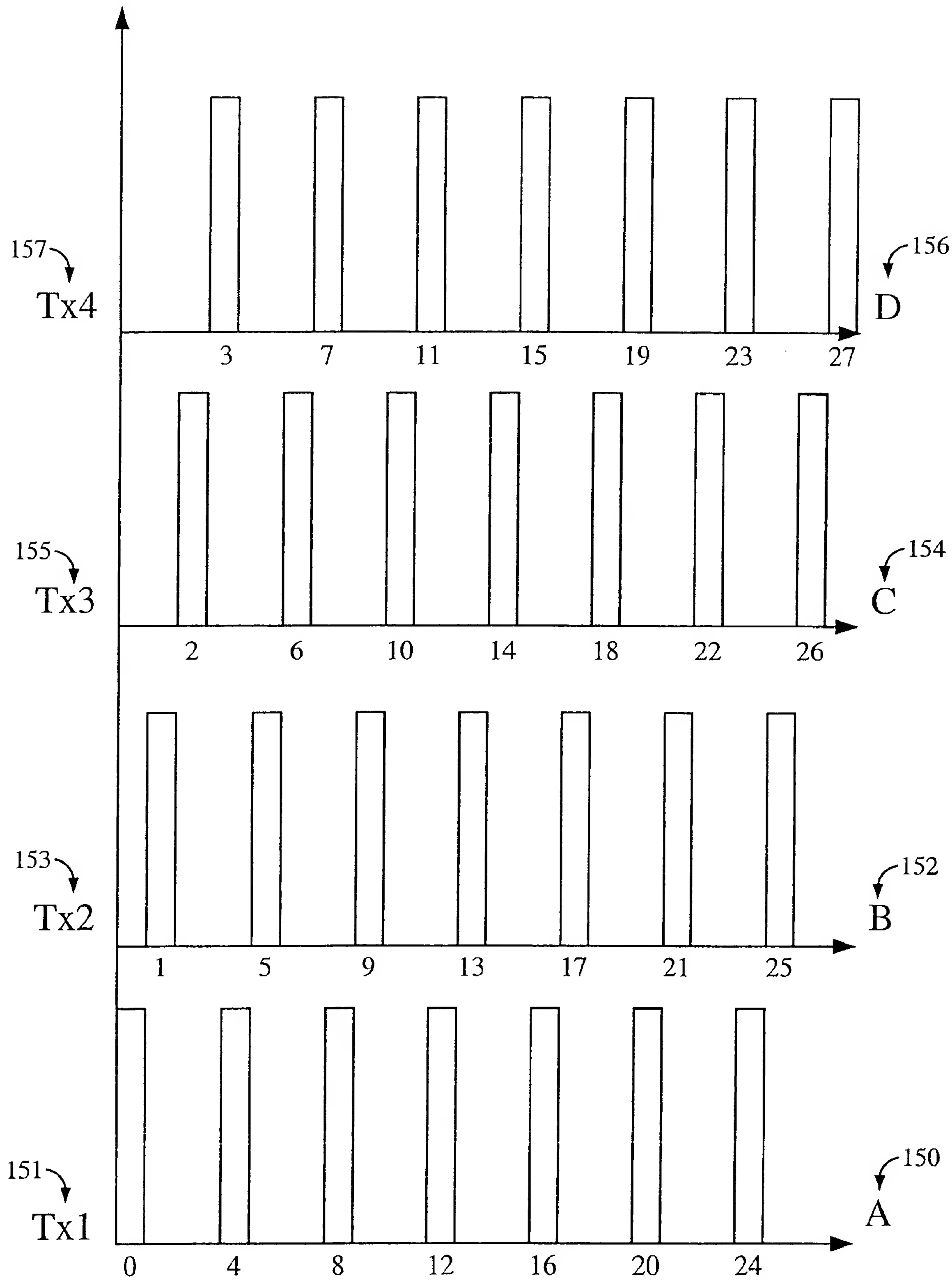

140
PRECONDITION DATA; AND ASSIGN PRECONDITIONED DATA TO MULTIPLE SUB-CHANNELS AND ANTENNAS
141
142
PERFORM IFFT ON DATA ASSIGNED TO EACH ANTENNA
143
APPEND CYCLIC PREFIX/EXTENSION
144
TRANSMIT FROM MULTIPLE ANTENNAS
149
CSI FEEDBACK
145
146
RECEIVE AT MULTIPLE ANTENNAS
147
PERFORM FFT
148
MEASURE CSI ON PILOT; AND DEMODULATE USING CSI

FIG. 1B



OFDM SUB-CHANNELS

FIG. 1C

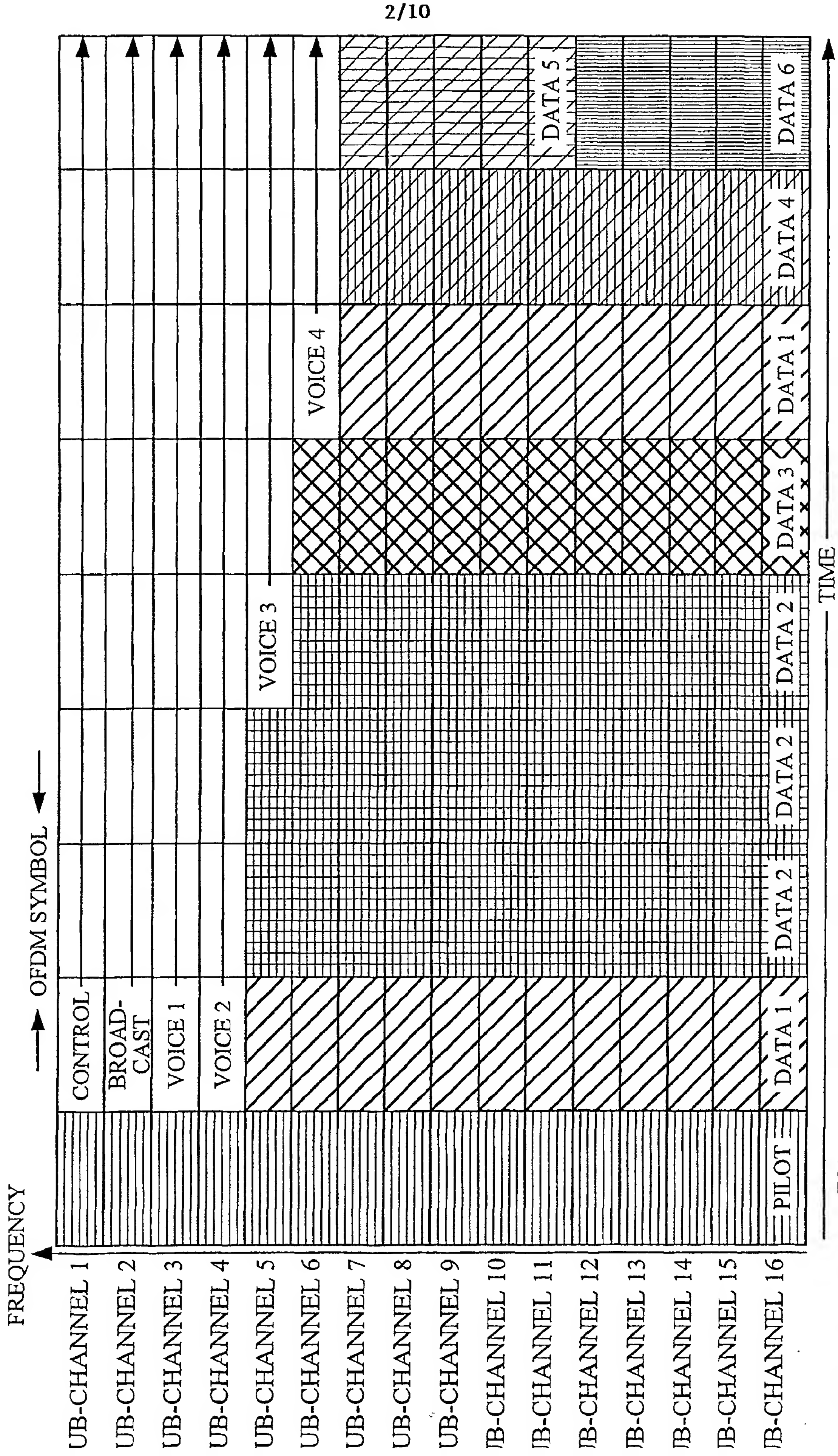

FREQUENCY
OFDM SYMBOL
TIME
UB-CHANNEL 1
UB-CHANNEL 2
UB-CHANNEL 3
UB-CHANNEL 4
UB-CHANNEL 5
UB-CHANNEL 6
UB-CHANNEL 7
UB-CHANNEL 8
UB-CHANNEL 9
JB-CHANNEL 10
JB-CHANNEL 11
JB-CHANNEL 12
JB-CHANNEL 13
JB-CHANNEL 14
JB-CHANNEL 15
JB-CHANNEL 16
PILOT
CONTROL
BROAD-CAST
VOICE 1
VOICE 2
VOICE 3
VOICE 4
DATA 1
DATA 2
DATA 2
DATA 2
DATA 3
DATA 1
DATA 4
DATA 5
DATA 6
TS1
TS2
TS3
TS4
TS5
TS6
TS7
TS8
TS9

FIG. 2

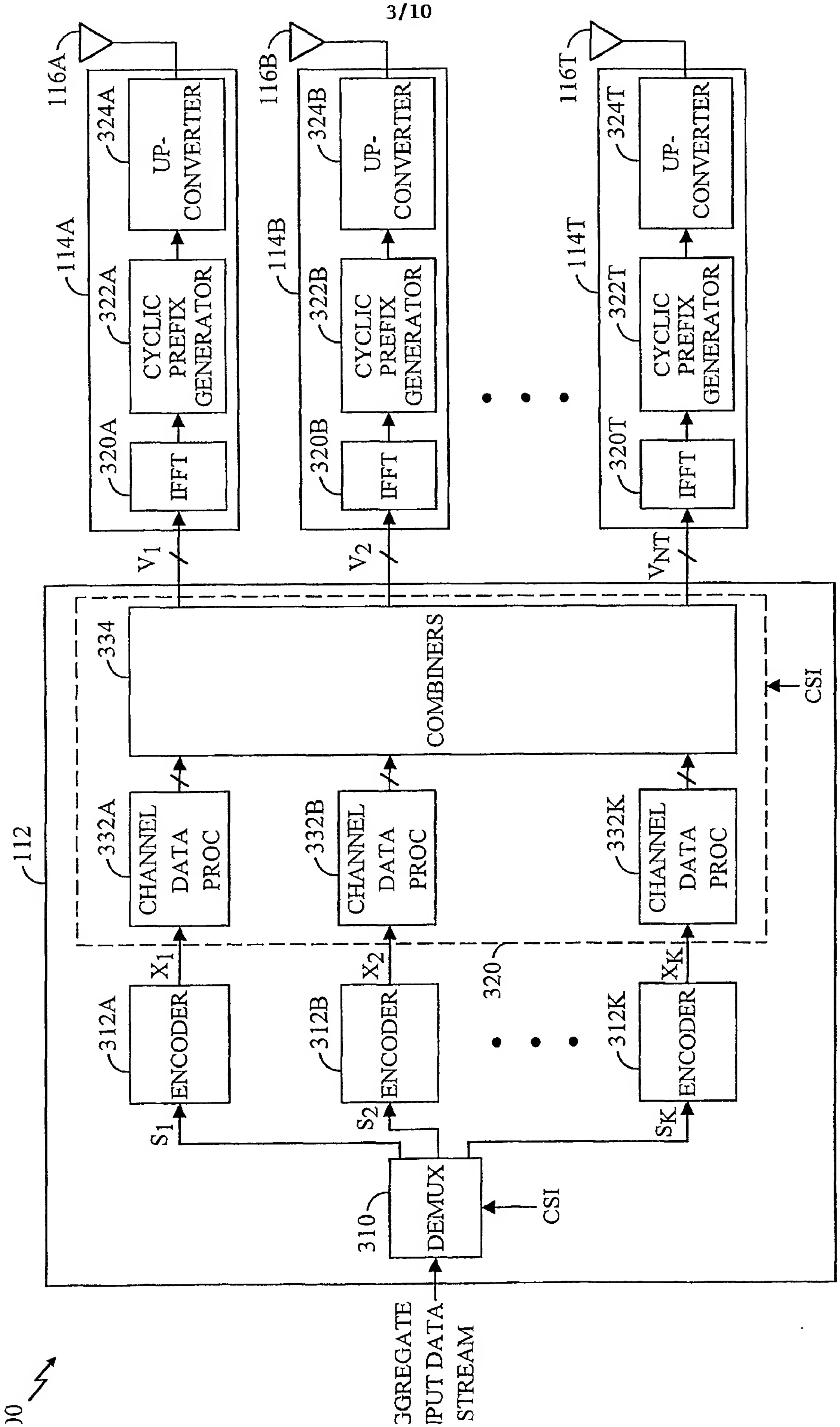

116A
116B
116T
114A
114B
114T
324A
324B
324T
UP-CONVERTER
322A
322B
322T
CYCLIC PREFIX GENERATOR
320A
320B
320T
IFFT
V_1
V_2
V_{NT}
334
COMBINERS
CSI
332A
332B
332K
CHANNEL DATA PROC
112
320
X_1
X_2
X_K
312A
312B
312K
ENCODER
S_1
S_2
S_K
310
DEMUX
CSI
GGREGATE
IPUT DATA
STREAM

FIG. 3

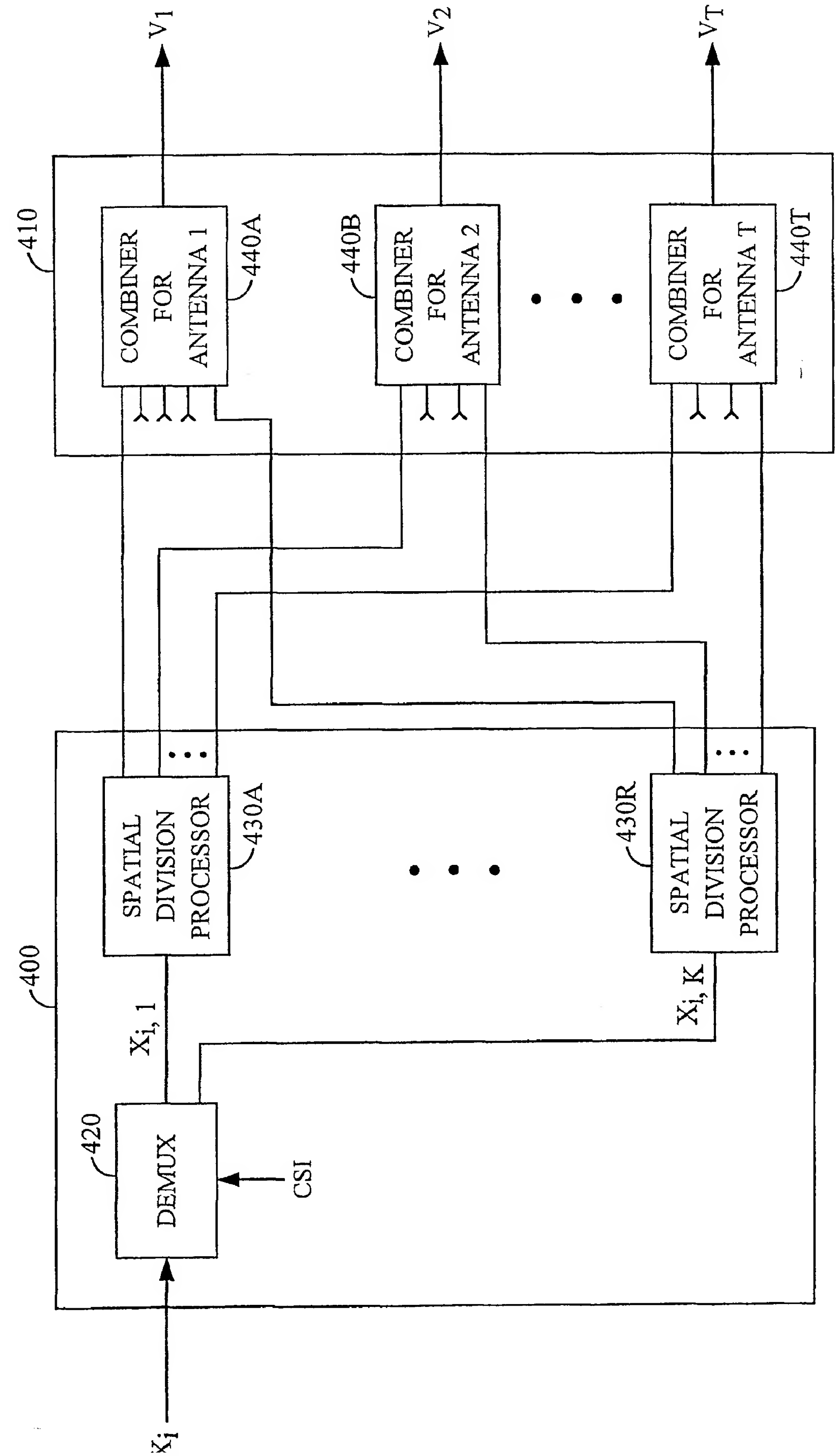

V_1
V_2
V_T
410
COMBINER FOR ANTENNA 1
440A
440B
COMBINER FOR ANTENNA 2
COMBINER FOR ANTENNA T
440T
SPATIAL DIVISION PROCESSOR
430A
430R
SPATIAL DIVISION PROCESSOR
400
$X_{i,1}$
$X_{i,K}$
420
DEMUX
CSI
X_i

FIG. 4A

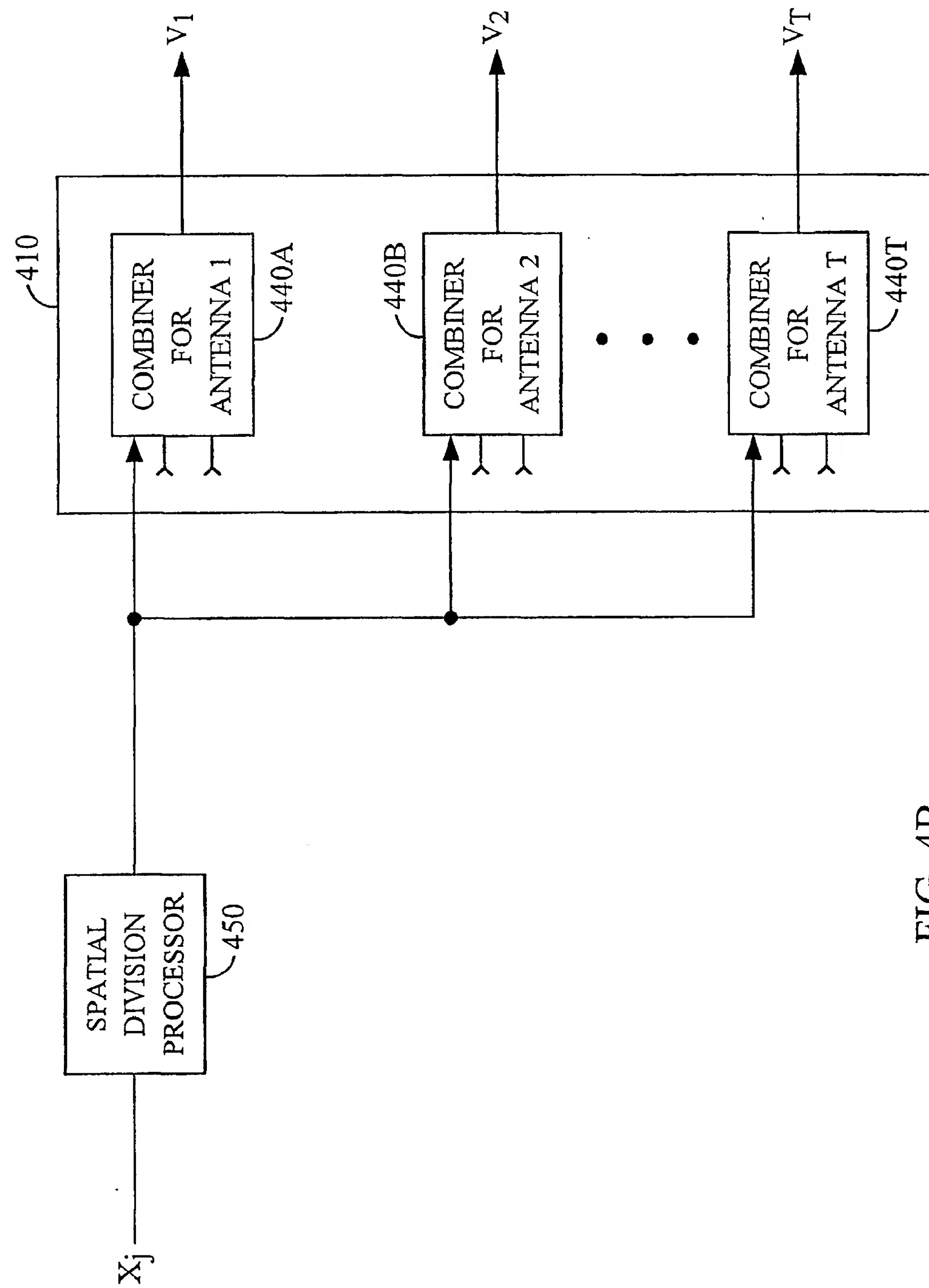

V_1
V_2
V_T
410
COMBINER FOR ANTENNA 1
440A
440B
COMBINER FOR ANTENNA 2
COMBINER FOR ANTENNA T
440T
SPATIAL DIVISION PROCESSOR
450
X_j

FIG. 4B

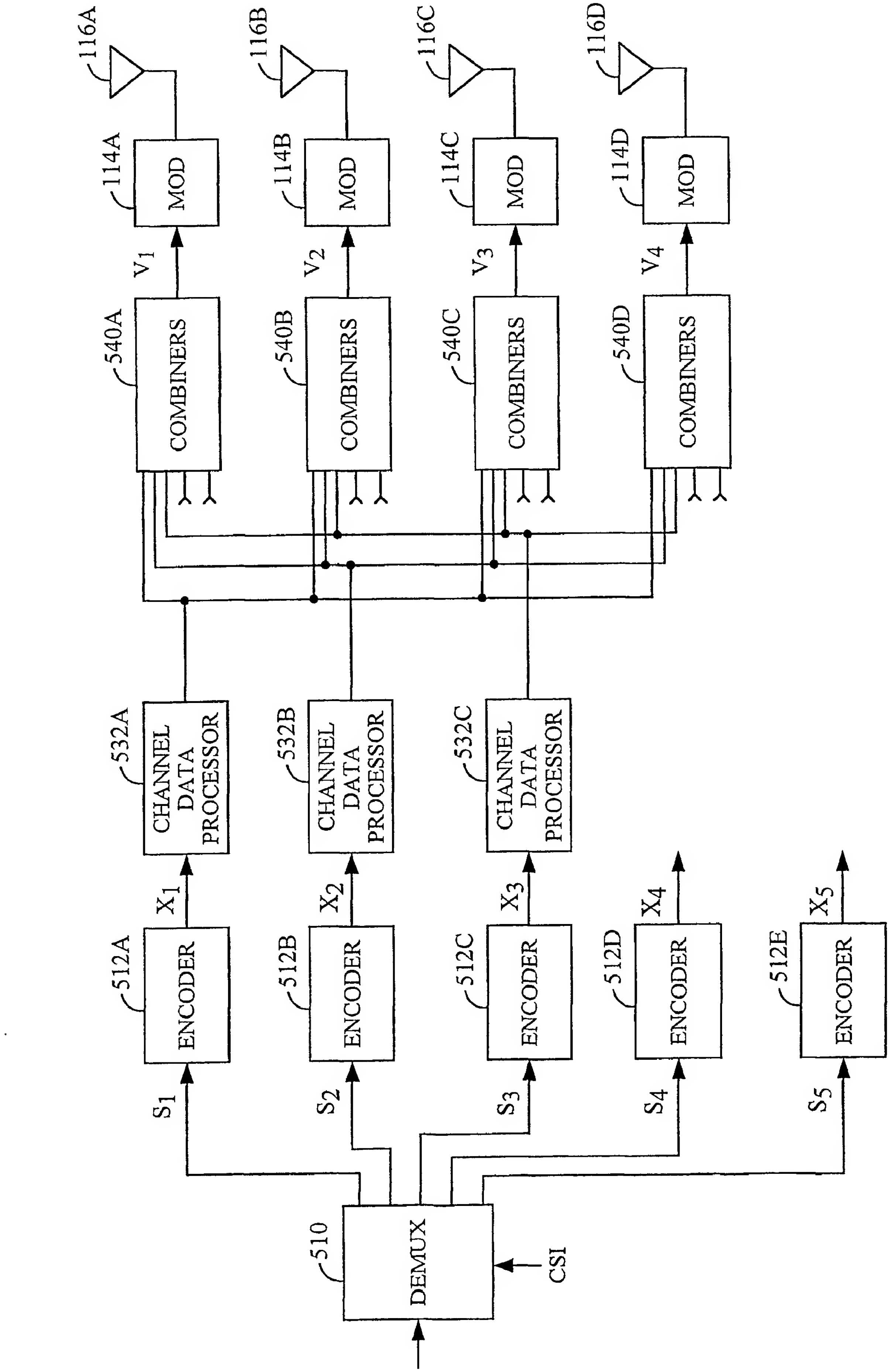

116A
116B
116C
116D
114A
MOD
114B
MOD
114C
MOD
114D
MOD
V_1
V_2
V_3
V_4
540A
COMBINERS
540B
COMBINERS
540C
COMBINERS
540D
COMBINERS
532A
CHANNEL DATA PROCESSOR
532B
CHANNEL DATA PROCESSOR
532C
CHANNEL DATA PROCESSOR
X_1
X_2
X_3
X_4
X_5
512A
ENCODER
512B
ENCODER
512C
ENCODER
512D
ENCODER
512E
ENCODER
S_1
S_2
S_3
S_4
S_5
510
DEMUX
CSI

FIG. 5A

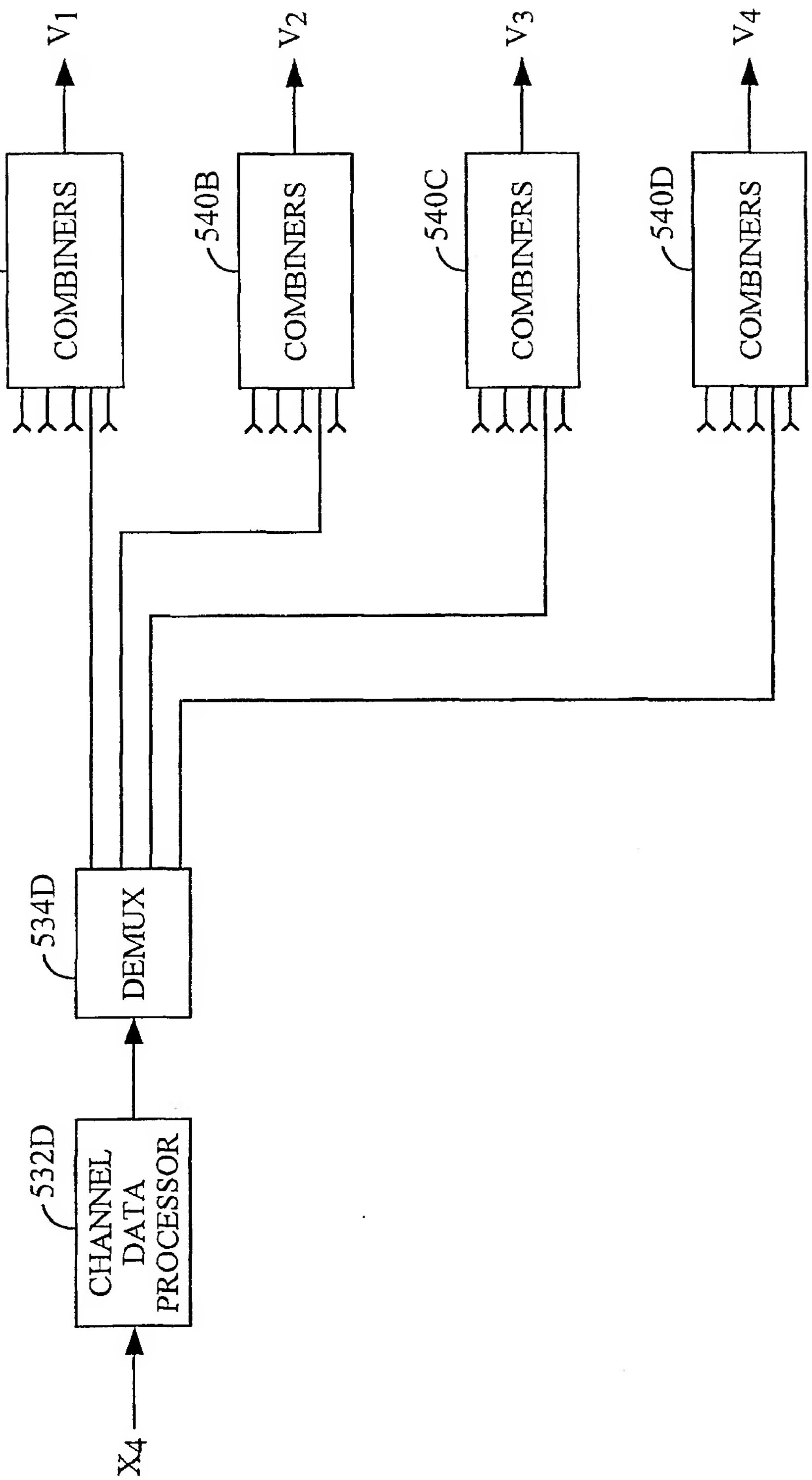

X_4
532D
CHANNEL DATA PROCESSOR
534D
DEMUX
540A
COMBINERS
540B
COMBINERS
540C
COMBINERS
540D
COMBINERS
V_1
V_2
V_3
V_4

FIG. 5B

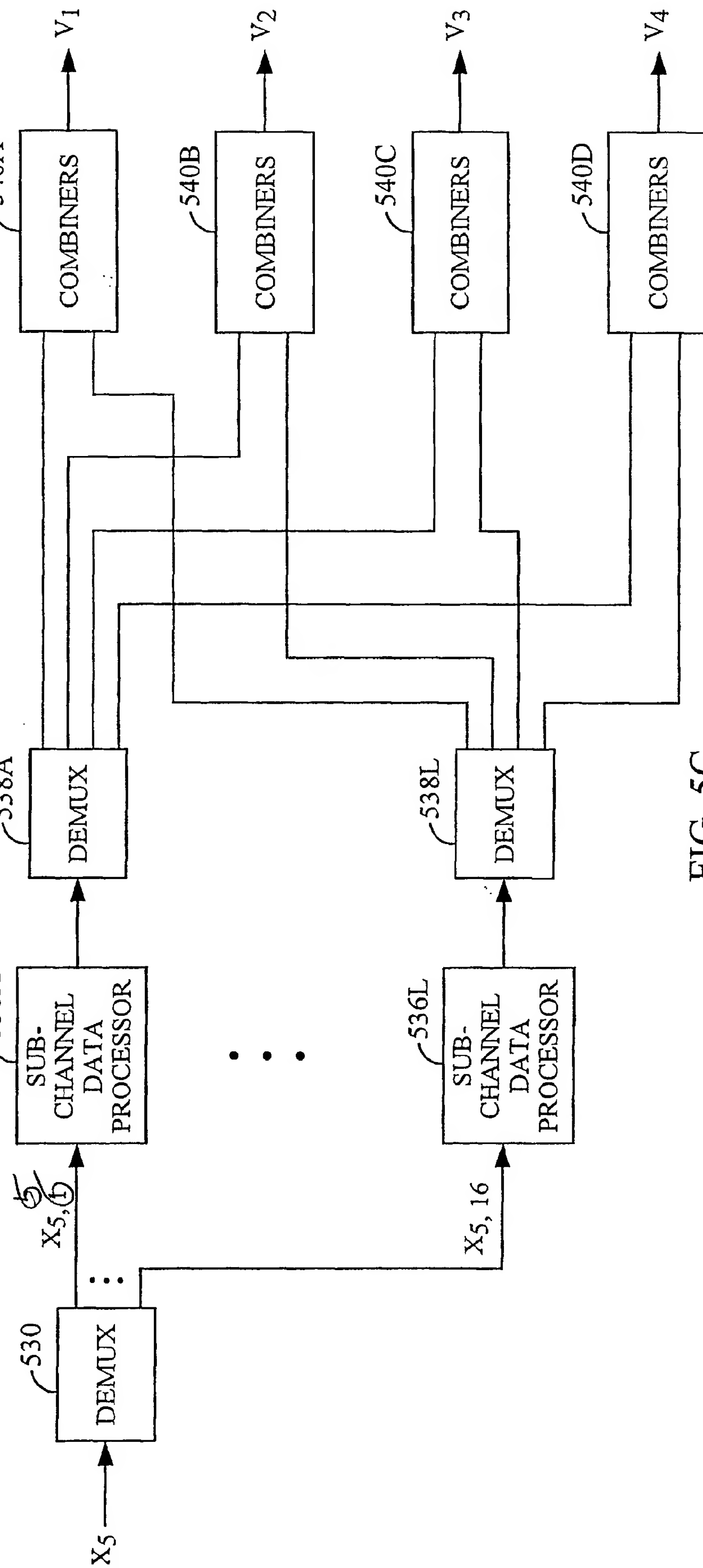

V_1
V_2
V_3
V_4
540A
COMBINERS
540B
COMBINERS
540C
COMBINERS
540D
COMBINERS
538A
DEMUX
538L
DEMUX
536A
SUB-CHANNEL DATA PROCESSOR
536L
SUB-CHANNEL DATA PROCESSOR
$X_{5,1}$
$X_{5,16}$
530
DEMUX
X_5

FIG. 5C

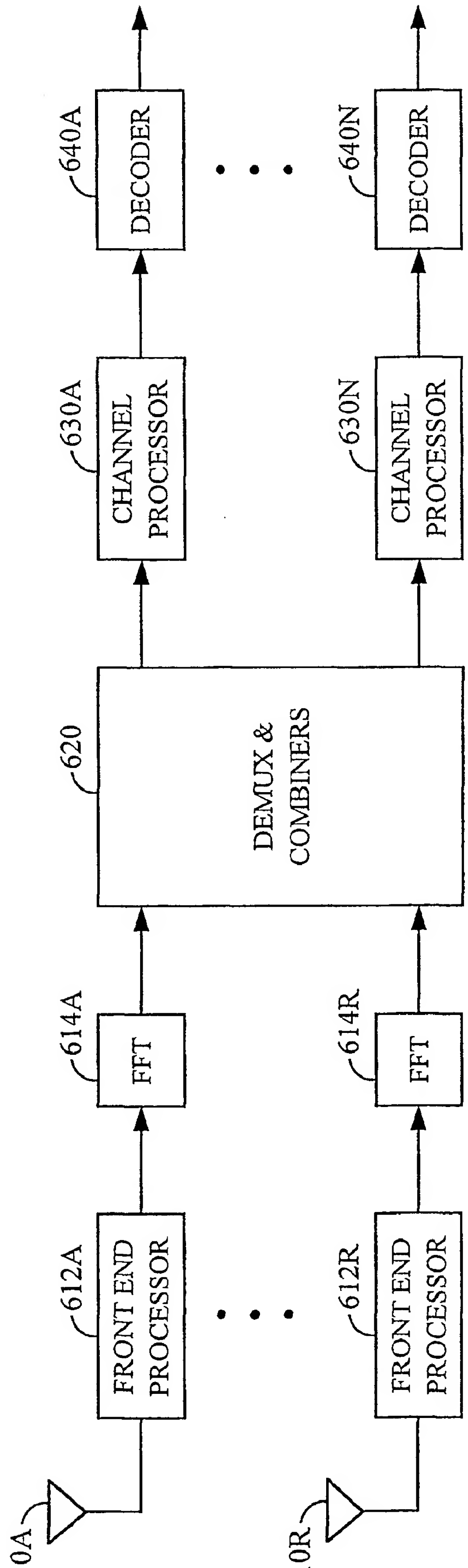

640A
DECODER
640N
DECODER
630A
CHANNEL PROCESSOR
630N
CHANNEL PROCESSOR
620
DEMUX & COMBINERS
614A
FFT
614R
FFT
612A
FRONT END PROCESSOR
612R
FRONT END PROCESSOR
0A
0R

FIG. 6

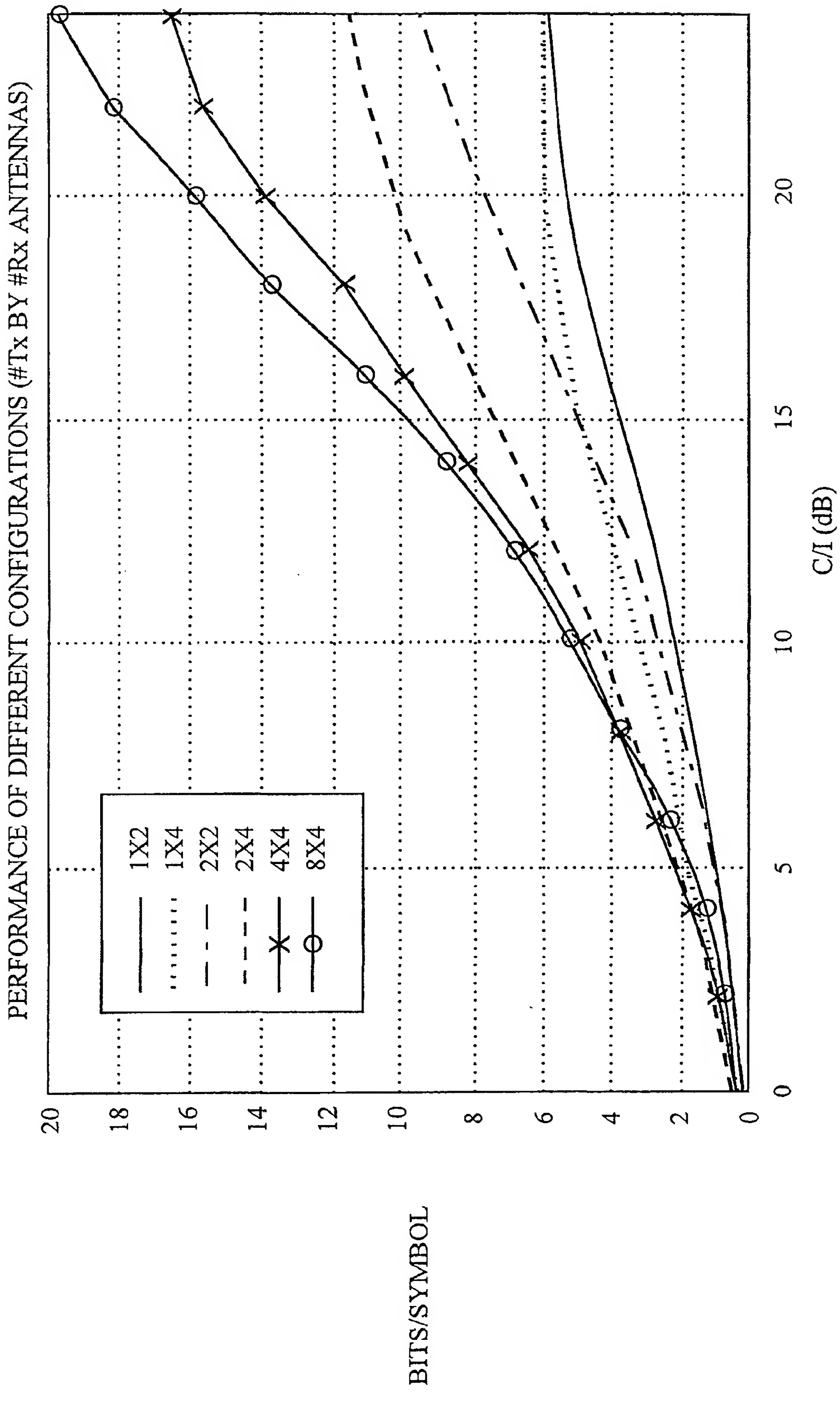
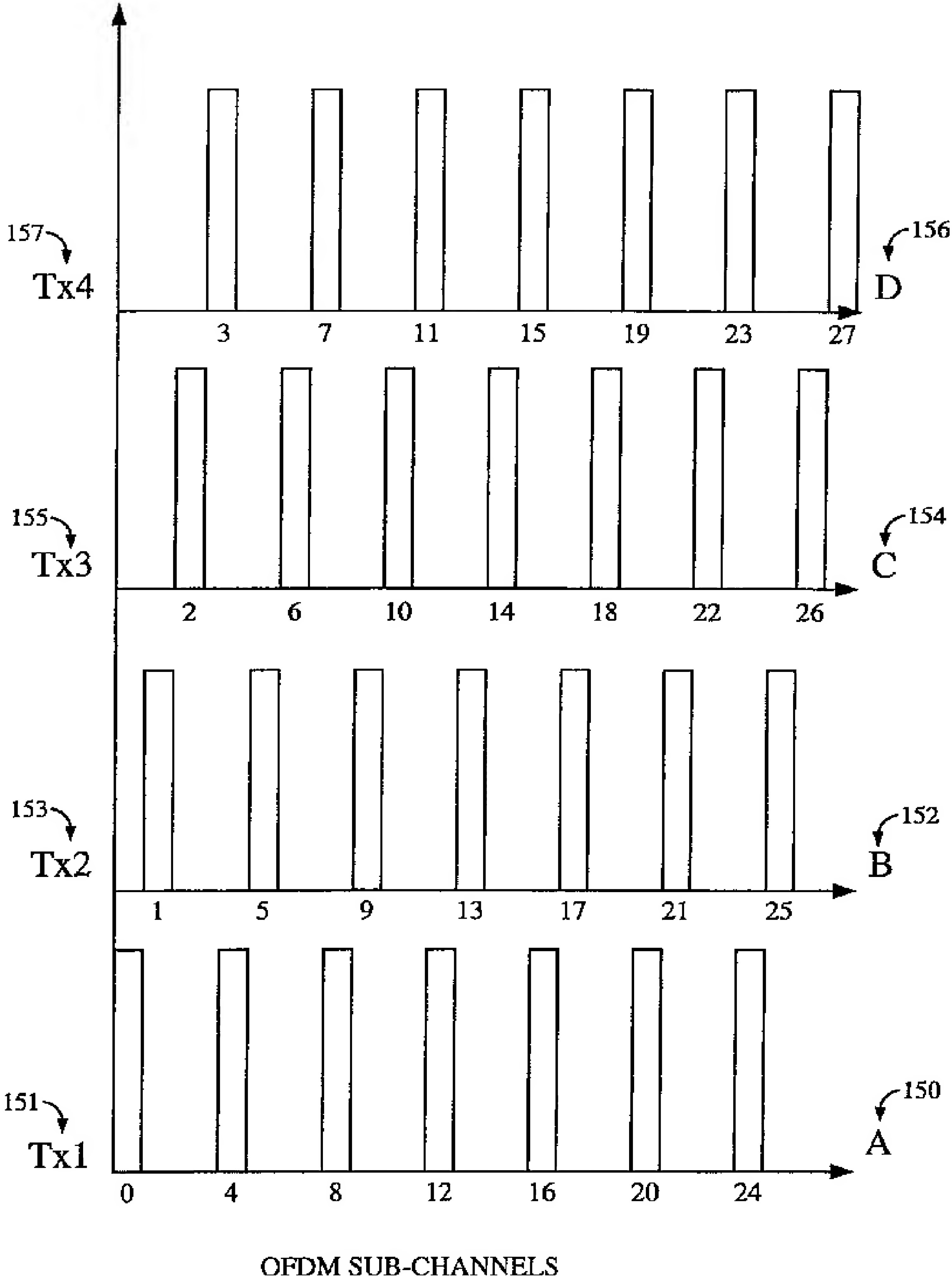

PERFORMANCE OF DIFFERENT CONFIGURATIONS (#Tx BY #Rx ANTENNAS)
1X2
1X4
2X2
2X4
4X4
8X4
BITS/SYMBOL
C/I (dB)
0
2
4
6
8
10
12
14
16
18
20
0
5
10
15
20

FIG. 7



157
Tx4
156
D
3
7
11
15
19
23
27
155
Tx3
154
C
2
6
10
14
18
22
26
153
Tx2
152
B
1
5
9
13
17
21
25
151
Tx1
150
A
0
4
8
12
16
20
24
OFDM SUB-CHANNELS

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1083 Bruxelles (BE)

(54) A method and apparatus for multi-user transmission

(57) The present invention is related to a method of transmitting data signals (50) from at least two transmitting terminals (20) with each at least one transmitting means (60) to at least one receiving terminal (40) with a spatial diversity receiving means (80) comprising the steps:

- transmitting from said transmitting terminals (20) transformed data signals (70), being transformed versions of said data signals; receiving on said spatial diversity means (80) received data signals being at least function of at least two of said transformed data signals (70);
- subband processing (90) of at least two of said received data signals in said receiving terminal (40); and
- determining estimates of said data signals (120) from subband processed received data signals (140) in said receiving terminal.

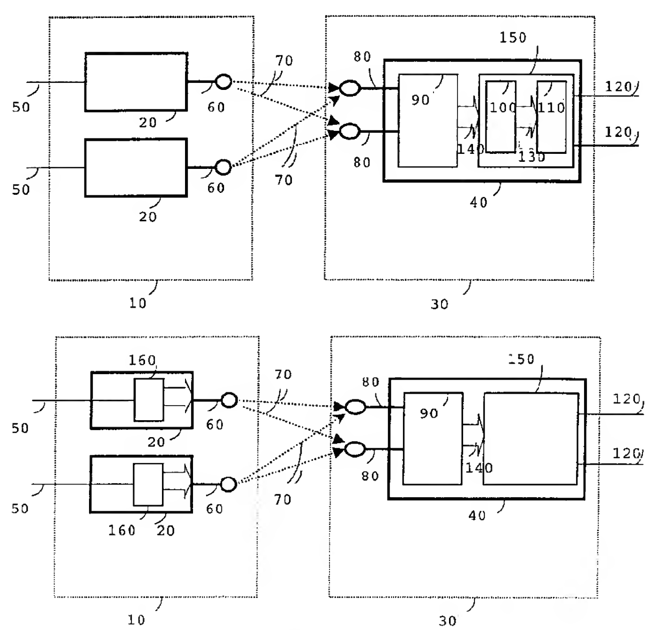

FIG. 1

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Description

Field of the invention

[0001] The present invention relates to an apparatus and methods for high-speed multi-user wireless communication.

State of the art

[0002] The idea to use multicarrier methods as a modulation technique is known in the art [Chang, R.W. "Synthesis of band-limited orthogonal signals for multichannel data transmission," Bell syst. Tech. J., vol.45, pp. 1775-1796, Dec. 1966], [Saltzberg, B.R. "Performance of an efficient parallel data transmission system", IEEE Trans. Comm. Technol., vol. COM-15, Dec. 1967]. Possible benefits following from multicarrier modulation have been mentioned in many articles [Meuller, T. Brueninghaus, K. and Rohling H. "Performance of Coherent OFDM-CDMA for Broadband Mobile Communications", Wireless Personal Communications 2, Kluwer Academic Publishers, 1996, pp. 295-305], [Kaiser, S. "OFDM-CDMA versus DS-CDMA: Performance Evaluation for Fading Channels", ICC '95, pp. 1722-1726]. Numerous theoretical publications have been written on said attractive modulation technique [Kalet, "The multitone Channel", IEEE Trans. Commun., vol. 37, no. 2, Feb. 1989], [Fazel, G. Fettweis, "Multi-Carrier Spread-Spectrum", Kluwer Academic Publ., 1997]. Specifically in multipath fading propagation situations, such as for example encountered in an indoor environment, multicarrier modulation is a beneficial technique. Indeed, thanks to the insertion of a guard interval containing a cyclic prefix, it enables a very efficient way of combatting ISI, being intersymbol interference. Moreover, adaptive loading techniques make it possible to considerably increase the throughput performances [L. Van der Perre, S. Thoen, P. Vandenameele, B. Gyselinckx, M. Engels. "Adaptive loading strategy for a high speed OFDM-based WLAN". In Globecom '98. Sydney, Australia, November 1998]. However, for a given carrier modulation, the bandwidth efficiency in terms of bits/sec/hertz is fixed. Given the massive growth of wireless communication and the importance of broadband services, the spectrum becomes increasingly scarce. One method to increase the capacity or the bandwidth efficiency of a wireless system, is to apply cellularization in order to reuse spectrum in different non-interfering cells. While this technique has been applied successfully in mobile telephone networks, it is -from an economic point of view-inappropriate for small- or medium-scale indoor networks as WLANs or home LANs. First of all, high operating frequencies (i.e. millimetre wave band) would be required to achieve a reasonable reuse factor [M. Chiani, D. Dardari, A. Zanella, O. Andrisano. " Service Availability of Broadband Wireless Networks for Indoor Multimedia at Millimeter Waves". In ISSSE '98. pp. 29-33, Pisa, Italy, September 1998],[T. Ithara, T. Manabe, M. Fujita, T. Matsui and Y. Sugimoto. "Research Activities on Millimeter-Wave Indoor Wireless Communications", in ICUPC '95, Tokyo, Japan, November 1995]. Secondly, cellularisation introduces an extra layer of hierarchy and complicates the protocol stack. Thirdly, cellularisation increases the installation effort. An alternative method that allows spectrum reuse and which has none of the disadvantages of cellularisation, is the application of Space Division Multiple Access (SDMA) techniques [A. Paulraj, C. Papadias. "Space-Time Processing for Wireless Communications", IEEE Signal Processing Magazine, pp. 49-83 , November 1997]. Making use of an antenna array, SDMA can separate different users communicating over the same frequency band and at the same time, by exploiting their distinct spatial signature. As such, it allows reuse within one cell of the cellularized space. SDMA has been proposed for single-carrier systems, where its benefits have been extensively proven [G. Tsoulos, M. Beach and J. MacGeehan, "Wireless personal communications for the 21st century: European technological advances in adaptive antennas", IEEE Communications Magazine, Vol. 35, No. 9, pp. 102-9, Sept 1997],[R. Roy, "An overview of smart antenna technology and its application to wireless communication systems", in IEEE International Conf. On Personal Wireless Communications, pp234-8, New York, NY, 1997],[S. Jeng, G. Xu, H. Lin and W. Vogel, "Experimental study of antenna arrays in indoor wireless applications", in Asilomar Conference on Signals, Systems and Computers, pp. 766-70 Los Alamitos, CA, 1996]. However, these single-carrier SDMA systems for high speed (e.g. 100Mbps) wireless systems demand a massive amount of processing (e. g. in the order of Gflops) [P. Vandenameele, L. Van der Perre, B. Gyselinckx, M. Engels and H. De Man, "An SDMA Algorithm for High-Speed Wireless LAN", in Globecom 98 Sydney, Australia, pp. 189-194, November 1998]. The combination of OFDM as a modulation technique with an antenna array is known in the art [G. Raleigh and J. Cioffi, "Spatio-Termporal Coding for Wireless Communication", IEEE Transaction on Communications, Vol. 46, No.3, pp. 357-366, March 1998]. However, these algorithms are limited to a single user scenario and do not enable SDMA.

Aims of the invention

[0003] The aim of the present invention is to increase the performance/cost ratio of high-speed wireless networks by providing communication methods (also denoted transmission methods, being transmitting from at least one peer and receiving on at least one other peer) and a dedicated apparatus being inherently multi-user, multi-carrier and exploiting space division multiple access principles. Said transmission methods and said apparatus enables high spec-

trum efficient communication under multipath fading conditions.

Summary of the invention

[0004] In a first aspect of the invention (Figure 1) a method of transmitting data signals from at least two transmitting terminals with each at least one transmitting means to at least one receiving terminal with a spatial diversity receiving means is disclosed. Said spatial diversity receiving means can be a plurality of at least two antenna's being spaced apart or having a different polarization. The transmitting terminals can be grouped in a composite peer. The receiving terminals can be grouped in a processing peer. The spatial diversity receiving means comprises of at least two receiving means related to each other in such a way that said receiving means provide different spatial samples of a same data signal. In a first step in said transmitting method from said transmitting terminals transmitted data signals are transmitted. This transmission can be substantially simultaneously. The spectra of said transmitted data signals can be at least partly overlapping. The transmitted data signals are transformed versions of said data signals. In a second step in said transmitting method on said spatial diversity means received data signals are received. Said received data signals are at least function of at least two of said transmitted data signals. In a third step at least two of said received data signals are subband processed in said receiving terminal. In a last step estimates of said data signals are determined from said subband processed received data signals in said receiving terminal.

[0005] Subband processing of a data signal having a data rate, comprises in principle of splitting said data signal in a plurality of data signals, with a lower data rate and modulating each of said plurality of data signals with another carrier. Said carriers are preferably orthogonal. In an embodiment said subband processing of a data signal can be realized by using serial-to-parallel convertors and using a transformation on a group of data samples of said data signal.

[0006] The transmission method exploits thus a multi-carrier approach but then in a multi-user context as at least two transmitting terminals are present. The transmission method separates different users or transmitting terminals based on the different spatial samples of the signals received on the spatial diversity means. As such the transmission method can be understood as being a Space Division Multiple Access technique but then in a multi-carrier constellation instead of single carrier. However the method is more than a straightforward concatenation of a Space Division Multiple Access technique and a multi-carrier method. Indeed such a concatenation would result in time-domain processing of the data signals while the invention exploits the inherent frequency parallelism of subband processing techniques which result in a low complexity processing of the data signals in the frequency domain, compared to plain SDMA, being a time-domain technique. This enables low complexity nonlinear processing of the data signals, which improves the performance considerably. Indeed by such nonlinear processing of the data signals the invention exploits the frequency diversity observed for the different users. It can be mentioned that the data signals to be transmitted are often at least partially independent as they are originating from different users, although the transmission method does not rely on this.

[0007] In an embodiment of this first aspect of the invention subband by subband processing is disclosed. Said subband by subband processing can also be denoted per-subband or per-carrier processing.

[0008] In another embodiment of this first aspect of the invention a method for successive interference cancellation is disclosed. Said successive interference cancellation can be but is not limited to be realized in a subband by subband processing approach.

[0009] In yet another embodiment of this first aspect of the invention a method for interference dependent state insertion is disclosed.

[0010] In still a further embodiment of this first aspect of the invention a method for exploiting coherence grouping (being subband grouping) during initialization is disclosed.

[0011] Said embodiments of said first aspect can be combined.

[0012] In a second aspect of the invention a method of transmitting data signals from at least one transmitting terminal with a spatial diversity transmitting means to at least two receiving terminals with at least one receiving means is disclosed. The transmitting terminals can be grouped in a processing peer. The receiving terminals can be grouped in a composite peer. The spatial diversity transmitting means comprises of at least two transmitting means related to each other in such a way that said transmitting means provide different spatial samples of a same data signal. In a first step of the transmitting method combined data signals are determined in said transmitting terminal. Said combined data signals are transformed versions of said data signals. In a second step of the invented transmitting method said combined data signals are inverse subband processed. In a next step said inverse subband processed combined data signals are transmitted with said spatial diversity means. The spectra of said transmitted inverse subband processed combined data signals can be at least partly overlapping. In a further step on at least one of said receiving means of at least one receiving terminal inverse subband processed received data signals are received and then estimates of said data signals are determined from said inverse subband processed received data signals. It can be mentioned that said data signals are often at least partially independent as their destinations are different users, but the invented transmission method does not rely on that. Said first and second aspect of the invention can be combined.

[0013] In a third aspect of the invention an apparatus for determining estimates of data signals from at least two at least substantially simultaneously received data signals is disclosed. Said received data signals have at least partly overlapping spectra. Said apparatus comprises at least of at least one spatial diversity receiving means, circuitry being adapted for receiving said received data signals with said spatial diversity receiving means, circuitry being adapted for subband processing at least two of said received data signals and circuitry being adapted for determining estimates of said data signals from subband processed received data signals. Said apparatus can be exploited in uplink transmission methods in the processing peer.

[0014] In an embodiment of this third aspect of the invention parallelism in the apparatus structure is disclosed. This kind of parallelism can be exploited due to the inherent frequency parallelism, typical for the invented transmission methods. Indeed said circuitry being adapted for determining estimates of said data signals from subband processed received data signals can comprises a plurality of circuits each being adapted for determining part of said estimates of said data signals based on part of the subbands of said subband processed received data signals.

[0015] In another embodiment of this third aspect of the invention further parallelism is introduced in the apparatus structure in the circuitry being adapted for receiving said received data signals.

[0016] In a fourth aspect of the invention an apparatus for transmitting inverse subband processed combined data signals comprising at least of at least one spatial diversity transmitting means, circuitry being adapted for combining data signals, circuitry being adapted for inverse subband processing combined data signals, and circuitry being adapted for transmitting inverse subband processed combined data signals with said spatial diversity means, is disclosed. Said apparatus can be exploited in the downlink transmission methods in the processing peer.

[0017] In an embodiment of the invention parallelism in the apparatus structure is introduced either in said circuitry being adapted for combining data or (and) in said circuitry being adapted for transmitting inverse subband processed combined data signals.

[0018] An apparatus having the functionality and architectural characteristics of both said apparatus in said third and fourth aspect of the invention can be defined.

Brief description of the drawings

[0019] Figure 1: (Up-link) communication setup with a composite peer (10), comprising of at least two transmitting terminals (20), each having at least one transmitting means (60) and a processing peer (30), comprising of at least one receiving terminal (40), having a spatial diversity receiving means (being represented here but not limited thereto by two spaced apart receiving means (80)). Said receiving terminal (40) at least performs subband processing (90). The top of figure 1 shows a concentrated scenario wherein said receiving terminal further performs inverse subband processing (110). The bottom of figure 1 shows a split scenario wherein the transmitting terminals further performs inverse subband processing (160).

[0020] Figure 2: (Down-link) communication setup with a composite peer (340), comprising of at least two receiving terminals (330), each having at least one receiving means (320) and a processing peer (230), having at least one transmitting terminal (240), having a spatial diversity transmitting means (being represented here but not limited hereto by two spaced apart transmitting means (220)). Said transmitting terminal (240) performs at least inverse subband processing (260). The top of figure 2 shows a concentrated scenario wherein said transmitting terminal (240) further performs subband processing (280). The bottom of figure 2 shows a split scenario wherein the receiving terminals performs subband processing (350).

[0021] Figure 3 shows the performance of an embodiment of the invention being uplink Least-Squares-OFDM/SDMA for one to four simultaenous users. The Bit Error Rate (BER) is shown as function of the Signal to Noise Ratio (SNR). Said embodiment can be exploited for an arbitrary number of users. In said embodiment OFDM is exploited as subband method, through exploiting of inverse fast Fourier transform and fast Fourier transform. Said determining of estimates of data signals from subband processed received data signals is in said embodiment based on Least-Squares methods.

[0022] Figure 4 shows the performance of pcSIC (per-carrier successive interference cancellation) for uplink OFDM/SDMA for one to four simultaneous users. The Bit Error Rate (BER) is shown as function of the Signal to Noise Ratio (SNR). Said embodiment can be exploited for an arbitrary number of users. Said determining of estimates of data signals from subband processed received data signals is in said embodiment based on determining an estimate of a selected data signal, in a Least-Squares sense, then modifying subband processed received data signals, and finally determining estimates of the remaining data signals, being all data signals except the selected data signal. For said remaining data signals said determining of estimates of a selected data signal can be exploited also. Said selection of data signals approach results in introducing an ordering in which estimates of data signals will be determined.

[0023] Figure 5: Error propagation with uplink SIC-OFDM/SDMA. The error count and the Signal Interference Ration (SIR) is shown as function of frequency. Said observation of the error propagation motivates the use of State Insertion methods.

[0024] Figure 6 shows the performance of pcSIC for uplink OFDM/SDMA with State Insertion for one to four simul-

taneous users. Said embodiment can be exploited for an arbitrary number of users. The Bit Error Rate (BER) is shown as function of the Signal to Noise Ration (SNR). Said determining of estimates of data signals from subband processed received data signals in said embodiment based on determining a plurality of estimates of a selected data signals, then accordingly modifying said subband processed received data signals, then determining estimates of at least one of the remaining data signals, being all the data signals except the selected data signal, and finally selecting one of said plurality of estimates of said selected data signal. Said selection of data signals approach results in introducing an ordering in which estimates of data signals will be determined. The amount of estimates to be determined per data signal can differ and can possibly be one for some data signals.

[0025] Figure 7 shows the performance of downlink OFDM/SDMA for one to four simultaneous users. Said embodiment can be exploited for an arbitrary number of users. The Bit Error Rate (BER) is shown as function of the Signal toNoise Ratio (SNR).

[0026] Figure 8 shows the performance of uplink OFDM/SDMA for different coherence grouping factors. The Bit Error Rate (BER) is shown as function of the Signal to Noise Ration (SNR). In said embodiment the initialization, being the determining of relations between data signals and subband processed received data signals, is performed on a set-by-set basis. Said coherence grouping factor denote the amount of subbands in a set.

[0027] Figure 9 presents the operation count (initialization and processing) of Least-Squares-OFDM/SDMA.

[0028] Figure 10 present the operation count (initialization and processing) of pcSIC (per-carrier successive interference cancellation)-OFDM/SDMA.

[0029] Figure 11 presents the additional operation count for State Insertion (initialization and processing).

Detailed description of the invention

[0030] The invention is not limited to the detailed description of the invention found below.

[0031] The invention concerns (wireless) communication between terminals (20) (40) (330) (240). We can logically group the terminals on each side of the communication and call them peers (340) (230) (10) (30). Each peer can embody one or more terminals, whereby at least one peer embodies more than one terminal. The focus of the invention is thus multi-user. The peers can be transmitting and/or receiving information. For example, the peers can communicate in a half-duplex, being either transmitting or receiving at one time instance or a full duplex fashion, being substantially simultaneously transmitting and receiving.

[0032] The invention introduces Space Division Multiple Access (SDMA) techniques for systems making use of subband processing, and thus fits in a multi-carrier approach. In the invention, at least one of the communication peers (30) (230) consists of terminal(s) (40) (240) disposing of transmitting and/or receiving means (80) (220) that are able to provide different spatial samples of the transmitted and/or received signals. We will call these transmission and/or receiving means spatial diversity means. We will call the peer(s) disposing of said spatial diversity means the processing peer(s) (30)(230). Said processing peer communicates with at least two terminals at the opposite peer (10) (340), which can operate at least partially simultaneously and the communicated signals' spectra can at least partially overlap. Note that Frequency Division Multiple Access techniques rely on signals spectra being non-overlapping while Time Division Multiple Access techniques rely on communicating signals in different time slots thus not simultaneously. We will call this opposite peer(s) consisting of at least two terminals (20) (330) using the same frequencies at the same time, the composite peer(s) (10)(340). The invention concerns (wireless) communication between terminals whereby at least the processing peer(s) (30) (230) disposes of subband processing means.

[0033] The communication between the composite peer and the processing peer can consist of uplink (Figure 1) and downlink (Figure 2) transmissions. With uplink transmission is meant a transmission whereby the composite peer transmits data signals and the processing peer receives data signals. With downlink transmission is meant a transmission whereby the processing peer transmits data signals and the composite peer receives data signals. The uplink and downlink transmissions can either be simultaneous (full duplex) (for example using different frequency bands), or they can operate in a time-duplex fashion (half duplex)(for example using the same frequency band), or any other configuration.

[0034] (Wireless) transmission of data or a digital signal from a transmitting to a receiving circuit requires digital to analog conversion in the transmission circuit and analog to digital conversion in the receiving circuit. In the further description it is assumed that the apparatus in the communication set-up have transmission and receiving means, also denoted front-end, incorporating these analog to digital and digital to analog conversion means including amplification or signal level gain control and realizing the conversion of the RF signal to the required baseband signal and vice versa. A front-end can comprise of amplifiers, filters and mixers (down converters). As such in the text all signals are represented as a sequence of samples (digital representation), thereby assuming that the above mentioned conversion also takes place. Said assumption does not limit the scope of the invention though. Communication of a data or a digital signal is thus symbolized as transmitting and receiving of a sequence of (discrete) samples. Prior to transmission, the information contained in the data signals can be fed to one or more carriers or pulse-trains by mapping said data signals

to symbols which consequently modulate the phase and/or amplitude of the carrier(s)or pulse-trains (e.g. using QAM or QPSK modulation). Said symbols belong to a finite set, which is called the transmitting alphabet. The signals resulting after performing modulation and/or front-end operations on said data signals, are called transformed data signals, to be transmitted further.

[0035] After reception by the receiving means, the information contained in the received signals is retrieved by transformation and estimation processes. These transformation and estimation processes can include but should not include demodulation, subband processing, decoding, equalization. After said estimation and transformation processes, received data signals are obtained consisting of symbols belonging to a finite set, which is called the receiving alphabet. The receiving alphabet is preferably equal to the transmitting alphabet.

[0036] The invention can further exploit methods and means for measuring the channel impulse responses between the transmission and/or reception means of the individual terminals at the composite peer on the one hand, and the spatial diversity means of the processing peer on the other hand. The channel impulse responses measurement can be either obtained on basis of an uplink transmission and/or on basis of a downlink transmission. The thus measured channel impulse responses can be used by the processing peer and/or composite peer in uplink transmissions and/or in downlink transmissions. The invention can further exploits methods for determining the signal power of received data signals and methods for determining the interference ratio of data signals.

[0037] The spatial diversity means ensures the reception or transmission of distinct spatial samples of the same signal. This set of distinct spatial samples of the same signal is called a spatial diversity sample. In an embodiment, spatial diversity means embody separate antennas. In this embodiment, the multiple antennas belonging to one terminal can be placed spatially apart (as in Figure 1 and 2), or they can use a different polarization. The multiple antennas belonging to one terminal are sometimes collectively called an antenna array. The invention is maximally efficient if the distinct samples of the spatial diversity sample are sufficiently uncorrelated. In an embodiment, the sufficiently uncorrelated samples are achieved by placing different antennas apart over a sufficiently large distance. For example, the distance between different antennas can be chosen to be half a wavelength of the carrier frequency at which the communication takes place. Spatial diversity samples are thus different from each other due to the different spatial trajectory from the transmitting means to their respective receiving means or vice versa. Alternatively said spatial diversity samples are different from each other due to the different polarization of their respective receiving or transmitting means.

[0038] The methods described rely on the fact that at least the processing peer performs subband processing, called SP in the sequel, in the uplink mode (Figure 1), and inverse subband processing, called ISP in the sequel, in the downlink mode (Figure 2). Furthermore, in the uplink mode ISP takes place either in the composite peer prior to transmission (see Figure 1 bottom) or in the processing peer after SP (see Figure 1 top). In the downlink mode, SP takes place either in the composite peer after reception (see Figure 2 bottom) or in the processing peer before ISP (see Figure 2 top). The scenarios where both ISP and SP are in either transmission direction carried out in the processing peer, are called concentrated scenarios. The remaining scenarios, i.e. where ISP and SP are carried out in different peers in either transmission direction, are called split scenarios.

[0039] The uplink transmission methods can be formalized as a first aspect of the invention being methods of data signals (50) from at least two transmitting terminals (20) with each at least one transmitting means (60) to at least one receiving terminal (40) with a spatial diversity receiving means (80) comprising the following steps: (first step) transmitting from said transmitting terminals (20) transformed data signals (70), being transformed versions of said data signals; (second step) receiving on said spatial diversity means (80) received data signals being at least function of at least two of said (transmitted) transformed data signals (70); (third step) subband processing (90) of at least two of said received data signals in said receiving terminal (40); and (fourth step) determining estimates of said data signals (120) from (the obtained) subband processed received data signals (140) in said receiving terminal.

[0040] Said transmitting of said transformed data signals can be substantially simultaneously. The spectra of said transformed data signals can be at least partly overlapping.

[0041] In the uplink split scenario said transformation of said data signals (50) to transformed data signals (70) comprises inverse subband processing (160). In the uplink concentrated scenario said determining (150) of estimates of said data signals from (the obtained) subband processed received data signals in said receiving terminal comprises the following steps: (first step) determining (100) intermediate estimates of said data signals (130) from said subband processed received data signals in said receiving terminal; (second step) obtaining said estimates of said data signals (120) by inverse subband processing (110) said intermediate estimates.

[0042] The downlink transmission method can be formalized as a second aspect of the invention being methods of transmitting data signals (200) from at least one transmitting terminal (240) with a spatial diversity transmitting means (220) to at least two receiving terminals (330)with at least one receiving means (320) comprising the following steps: (first step) determining (250) combined data signals (300) in said transmitting terminal, said combined data signals being transformed versions of said data signals; (second step) inverse subband processing (260) said combined data signals; (third step) transmitting with said spatial diversity means (220) (the obtained) inverse subband processed

combined data signals; (fourth step) receiving on at least one of said receiving means (320) of at least one receiving terminal (330) inverse subband processed received data signals; (fifth step) determining estimates of said data signals from said inverse subband processed received data signals.

[0043] Said transmitting can be substantially simultaneously. The spectra of said (transmitted) inverse subband processed combined data signals can be at least partly overlapping;

[0044] In the downlink split scenario said determining of said estimates of said data signals in said receiving terminals comprises subband processing (350). In the downlink concentrated scenario determining combined data signals in said transmitting terminal comprises the following steps: (first step) determining intermediate combined data signals (290) by subband processing (280) said data signals; (second step) determining (270) said combined data signals from said intermediate combined data signals.

[0045] It must be mentioned that said transmission methods intend to transmit data signals from one peer to another peer but that due to transmission conditions in fact only estimates of said data signals can be obtained in the receiving peer. Said transmission methods naturally intend to be such that said estimates of said data signals are approximating said data signals as close as technically possible.

[0046] It is a characteristic of the invention that said transmission methods are not a straightforward concatenation of a Space Division Multiple Access technique and a multi-carrier modulation method. Indeed a straightforward concatenation would be to modulate in the transmitting terminal with such a multi-carrier method, then transmit in a Space Division Multiple Access setting and at the receiving terminal combine the different signals received on the spatial diversity means into a combined signal and then demodulate the combined signal. In the split scenario of the invention in the transmitting terminal the (to be transmitted) transformed signal is also modulated with a multi-carrier method and there is a transmission in a Space Division Multiple Access setting but in the receiving terminal first the different received signals on the spatial diversity means are demodulated and only then the demodulated received signals are combined. In the concentrated scenario both modulation and demodulating is concentrated in the receiving peer in the uplink and in the transmitting peer in the downlink. Note that here with modulation and demodulation subband and inverse subband processing is meant and not standard modulation.

[0047] In an embodiment of both the first and second aspect of the invention implements a multicarrier modulation technique. An example of such a multicarrier modulation technique uses IFFT as ISP and FFT as SP, and the modulation technique is called Orthogonal Frequency Multiplexing (OFDM) modulation. It can be stated that in said uplink transmission method that said subband processing is orthogonal frequency division demultiplexing. It can also be stated that in said uplink transmission method that said inverse subband processing is an orthogonal frequency division multiplexing. It can also be stated that in said downlink transmission method that said subband processing is orthogonal frequency division demultiplexing. It can also be stated that in said downlink transmission method that said inverse subband processing is orthogonal frequency division multiplexing.

[0048] In concentrated scenarios, the processing that is carried out in the processing peer on samples between SP (90) (280) and ISP (110) (260) is called subband domain processing (270)(100). In split scenarios, the processing that is carried out prior to ISP (160) (260) in the transmitting terminal(s) and after SP (90) (350) in the receiving terminal (s), is called subband domain processing (eg. (250)). With before is meant coming earlier in time during the transmission or the reception, and with after is meant coming later in time during the transmission or the reception. In concentrated scenarios, the signals (130) (140) (290) (300) between the SP and the ISP are called signals in subband domain representation. In split scenarios, the signals (50) (300) (200) before the ISP in the transmitting terminal(s) and the signals (360)(140)(120) after the SP in the receiving terminal(s) are called signals in a subband domain representation.

[0049] In an embodiment, the subband processing consists of Fast Fourier Transform (FFT) processing and the inverse subband processing consists of Inverse Fast Fourier Transform Processing. By FFT processing is meant taking the Fast Fourier Transform of a signal. By Inverse FFT processing is meant taking the Inverse Fast Fourier Transform of a signal.

[0050] In the invention the transmitted sequence is divided in data subsequences prior to transmission. Said data subsequences correspond to subsequences that will be processed as one block by the subband processing means. In case of multipath conditions, a guard interval containing a cyclic prefix or postfix is inserted between each pair of data subsequences in the transmitting terminal(s). If multipath propagation conditions are experienced in the wireless communication resulting in the reception of non-negligible echoes of the transmitted signal and the subband processing means consist of (an) FFT and/or IFFT operation(s), this guard introduction results in the equivalence between convolution of the time-domain data signals and the time-domain channel response with multiplication of the frequency-domain data-signals and the frequency-domain channel response. The insertion of said guard intervals can occur in both concentrated and split scenarios. One can thus state that in an embodiment of the invention in a split scenario, the transmitting terminal(s) insert guard intervals containing a cyclic prefix or postfix between each pair of data subsequences after performing ISP on the data subsequences and before transmitting the data subsequences. In another embodiment of the invention in a concentrated scenario, said guard intervals are inserted in the transmitted sequence between each pair of data subsequences without performing ISP on said data subblocks in the transmitting terminal

(s). This can be formalized as follows by stating that in said uplink transmission methods said transformation of said data signals to transmitted data signals further comprises of guard interval introduction. Said guard interval introduction can equally well be applied in the downlink transmission methods. Alternatively overlap and save techniques can be exploited also.

[0051] The terminal(s) disposing of the spatial diversity means (thus in the processing peer) dispose(s) of SP and/ or ISP means that enable subband processing of the distinct samples of the spatial diversity sample. Also, it disposes of means for combinatory processing. By combinatory processing means is meant means that process data coming from subbands of the distinct samples in the spatial diversity sample. In said combinatory processing means, different techniques can be applied to retrieve or estimate the data coming from the different distinct terminals or to combine the data to be transmitted to distinct terminals. The invention discloses methods for performing said combinatory processing, both for uplink transmission and for downlink transmission.

[0052] Combinatory processing in the uplink relates to a communication situation whereby the peer disposing of spatial diversity means, which is called the processing peer, is receiving signals from the composite peer, which embodies different terminals transmitting (at least partially simultaneous) transformed data signals (having at least partially overlapping spectra). With said determining of estimates of said data signals (120) from said subband processed received data signals (140) in said receiving terminal in said uplink transmission method is meant said combinatory processing.

[0053] Combinatory processing in the downlink relates to a communication situation whereby the peer disposing of spatial diversity means, which is called the processing peer, is transmitting signals to the composite peer, which embodies different terminals transmitting (at least partially simultaneous) so-called inverse subband processed combined data signals (having at least partially overlapping spectra). With determining (250) combined data signals (300) in said transmitting terminal in said downlink transmission method is meant said combinatory processing.

[0054] In the invention further the following notations are used: x is used for a transmitted data sample. The notation y is used for a received data sample. The notation n is used for a noise sample. The notation X is used for a transmitted data sample matrix. The notation Y is used for a received data sample matrix. The notation N is used for a noise sample matrix. The notation greek symbol sigma is used for the variance of the noise. The notation h(t) is used for the channel impulse response represented in the time domain. The notation h[s] is used for the channel impulse response represented in the frequency domain. The array index s (going from 1 to S) refers to the specific subband to which a sample or a channel impulse response corresponds. The notation S is used for the total number of subbands that are processed by the subband processing means. The superscript index u (going from 1 to U) refers to the individual terminal of the composite peer by which the data signal was sent in the uplink mode, or for which the data signal is intended in the downlink mode. The notation U is used for the number of simultaneous terminals of the same subbands in the composite peer. The subscript index a (going from 1 to A) refers to one specific spatial sample of the spatial diversity sample in the processing peer. The notation A is used for the number of distinct samples in the spatial diversity sample in the processing peer The notation e is used for an equalizer coefficient. The notation E is used for an equalizer coefficient matrix. The notation greek epsilon is used for the stochastical expectation operator. A ~ on top of a symbol indicates a soft estimation of the symbol. With a soft estimation is meant an estimation that is not necessarily contained in the receiving alphabet. A bar on top of a symbol indicates a hard estimation of the symbol. With a hard estimation is meant an estimation that is equal to a symbol contained in the receiving alphabet.

[0055] In the invention decision methods are exploited. Said decision methods obtain one or more intermediate hard estimates, on basis of a single soft estimate. In an embodiment of the invention, said decision methods obtain one intermediate hard estimate by determining that signal from the receiving alphabet that has the smallest distance to the soft estimate. In another embodiment of the invention, said decision methods obtain multiple intermediate hard estimates by determining those signals from the receiving alphabet that have the smallest distance to the soft estimate.

[0056] In the invention selection methods are exploited. Said selection methods obtain hard estimates for a specific data symbol on basis of intermediate hard estimates of said specific data symbol. In an embodiment of the invention, there is only one intermediate hard estimate for a specific data signal/symbol and the hard estimate is equal to the intermediate hard estimate. In another embodiment, there are several intermediate hard estimates for at least one data signal. From multiple intermediate hard estimates, one intermediate hard estimate for one specific data signal is selected as the hard estimate on basis of a probability criterion.

[0057] In the invention recombiner methods are exploited. Said recombiner methods obtain recombined spatial diversity samples, on basis of a hard or intermediate hard estimate, by calculating the spatial diversity samples that would have been received if the data symbol corresponding to said hard or intermediate hard estimate would have been transmitted.

[0058] In the invention modifier methods are exploited. Said modifier methods obtain modified spatial diversity samples, by applying the following steps. First, they obtain recombined spatial diversity samples by applying recombiner methods based on previously obtained hard or intermediate hard estimates. Secondly, they obtain modified spatial diversity samples by exploiting the recombined spatial diversity samples and the original spatial diversity samples.

[0059] Said modifier methods are exploited in the uplink transmission methods, denoted successive interference cancellation methods, wherein said determining of said estimates of said data signals from said subband processed received data signals in said receiving terminal further comprising for at least one data signal the following steps: (first step) selecting from said data signals a selected data signal; (second step) determining an estimate of said selected data signal from said subband processed received data signals; (third step) modifying said subband processed received data signals based on said estimate of said selected data signal via a modifier method; and (fourth step) determining estimates of said remaining data signals from said modified subband processed received data signals. Note that said selection of a selected data signals is just determining for which signal said method will be applied. Said selection should not be confused with the selection methods described above.

[0060] Said modifier methods are also exploited in the uplink transmission methods, denoted state insertion methods, wherein said determining of said estimates of said data signals from said subband processed received data signals in said receiving terminal further comprising for at least one data signal the following steps: (first step)selecting from said data signals a selected data signal; (second step) determining a plurality of estimates of said selected data signal from said subband processed received data signals; (third step) determining a plurality of modified subband processed received data signals, each of said modified subband processed received data signals being based on one of said estimates of said selected data signal, via a modifier method; (fourth step) determining a plurality of estimates of at least one of said remaining data signals from said plurality of modified subband processed received data signals; and (fifth step) thereafter selecting one of said estimates of said selected data signal (by a selection method). Note that said selection of a selected data signals is just determining for which signal said method will be applied. Said selection should not be confused with the selection methods described above.

[0061] It can be mentioned that said estimate of said selected data signal in said successive interference cancellation methods can be considered to be a hard estimate. Said plurality of said estimates of said selected data signal in said state insertion methods can be considered as intermediate hard estimates. Said plurality of estimates of said remaining data signals in both said methods can be either (intermediate) hard or soft estimates.

[0062] In the invention, methods for coherence grouping of subbands is presented. Said coherence grouping methods reduce the initialization effort, and can be used both in uplink and downlink transmissions. Said coherence grouping methods partition the S subbands in groups of adjacent subbands, each group i consisting of G_l subbands. The initialization computations are in said coherence grouping methods performed only once for each subgroup, instead of for each subband separately. Said coherence grouping methods do not affect the performances of the combinatory processing methods, provided all subbands in a subgroup experience sufficiently correlated channel impulse responses. Therefore, the numbers G_l are limited by the communication situation in order for the invention to be maximally efficient. For the initialization effort, said coherence grouping method results in a reduction of the computation complexity with a factor G, whereby G is the average of the numbers G_l of the subbands in the subgroups. In an embodiment, the communication is based on OFDM transmission and the numbers G_l are all chosen equal to a fixed part of the coherence bandwidth divided by the spacing between the carriers. The coherence bandwidth of the channel is the bandwidth over which the channel response is correlated. On a multipath propagation channel, said coherence bandwidth is inversely proportional to the relative delay of the echoes on the channel. In another embodiment, the numbers G_l are calculated on basis of a channel impulse response measurement, more specifically from the gradient of this channel impulse response. More formalized one can state that said determining of said estimates of said data signals in said receiving terminal comprises two steps. The first step is the initialization step wherein relations between said data signals and subband processed received data signals are determined. In the second step being the actual combinatory processing said relations between said data signals and said subband processed received data signals are exploited for determining said data signals. The coherence grouping method is then characterized by stating said subbands are grouped into sets, at least one set comprising of at least two subbands and that said initialization step is performed on a set-by-set basis.

[0063] In the invention, methods for combinatory processing for uplink communication are presented. Said combinatory processing for uplink communication methods obtain in the receiving terminal estimates for datasignals transmitted from one or more terminals in the composite peer, on basis of the subband processed spatial diversity samples, which can also be denoted subband processed received data signals.

[0064] In the invention, methods for per-subband combinatory processing for uplink communication are presented. Said per-subband combinatory processing for uplink communication methods obtain in the receiving terminal estimates for data signal(s) transmitted from one or more terminals in the composite peer and in one specific subband, on basis of the subband processed spatial diversity samples in that one specific subband, which can also be denoted subband processed received data signals in that one specific subband. Said per-subband combinatory processing methods can be formalized as methods in which said determining of estimates of said data signals in said receiving terminal is performed on a subband by subband basis.

[0065] In an embodiment of the invention, said methods for per-subband combinatory processing for uplink communication obtain estimates for data signal(s) from at least one terminal in the composite peer and in one specific subband,

by applying the following steps. First, they obtain soft estimates for the data signal(s) from each of these terminals and in that specific subband, by applying per-subband scalar combinatory processing for uplink communication methods. Secondly, they obtain estimates by applying decision methods on each of said soft estimates.

[0066] In the invention, methods for per-subband scalar combinatory processing for uplink communication are presented. Said per-subband scalar combinatory processing for uplink communication methods obtain soft estimates for data signal(s) in one subband and from one terminal in the composite peer, on basis of spatial diversity samples or modified spatial diversity samples in that subband. In an embodiment of the invention, said methods for per subband scalar combinatory processing for uplink communication obtain soft estimates for data signal(s) in one subband and from one terminal in the composite peer with linear methods. In this embodiment, the estimates of the data signal(s) transmitted by the terminal(s) of the composite peer $\tilde{x}^U$ [s] are calculated by linearly combining the single corresponding carrier signals or subbands received on the different antennas with the equalizer coefficients E[s], following Formula 1 below.

$$\underbrace{\begin{bmatrix} \tilde{x}^1[s] \\ \vdots \\ \tilde{x}^U[s] \end{bmatrix}}_{\tilde{X}[s]} = \underbrace{\begin{bmatrix} e_1^1[s] \dots e_A^1[s] \\ \vdots \qquad \vdots \\ e_1^U[s] \dots e_A^U[s] \end{bmatrix}}_{E[s]} \cdot \begin{bmatrix} y_1[s] \\ \vdots \\ y_A[s] \end{bmatrix} \qquad \text{Formula 1}$$

[0067] In an embodiment, the linear estimation is performed on basis of least-squares (LS) methods. In this embodiment of the invention, said E[s] is calculated to minimize the expectations given in Formula 2. For a given noise energy (sigma squared) and conditions on x given by Formula 3, E[s] obeys the U sets of linear equations of Formula 4, wherein the superscript H denotes the Hermitian transpose.

$$\mathcal{E}\left\{\left(x^u[s]-\tilde{x}^u[s]\right)*\left(x^u[s]-\tilde{x}^u[s]\right)\right\} \qquad \text{Formula 2}$$

$$\mathcal{E}\left\{x^u[s]*\tilde{x}^u[s]\right\}=1 \qquad \text{Formula 3}$$

$$\left[H[s]H[s]^H + \sigma^2 I_{A\times A}\right]E^H - H[s]=0 \qquad \text{Formula 4}$$

[0068] In an embodiment of the invention, said methods for per subband scalar combinatory processing for uplink communication obtain soft estimates for data signal(s) in one subband and from one terminal in the composite peer by linear zero forcing (ZF) methods. In this embodiment of the invention, said E[s] is calculated to maximally annihilate the channel distortion without taking the noise energy into account, in a so-called zero-forcing way. In this embodiment E[s] obeys the U sets of linear equations of Formula 5.

$$\left[H[s]^H\right]E^H - I_{U\times U}=0 \qquad \text{Formula 5}$$

[0069] In an embodiment of the invention, said methods for per subband combinatory processing for uplink commu-

nication obtain soft estimates for data signal(s) in one subband and from one terminal in the composite peer with non-linear methods such as for example maximum likelihood symbol estimation (MLSE).

[0070] In the invention, methods for combinatory processing for downlink communication are presented, called downlink combinatory processing methods. Said downlink combinatory processing methods are carried out in the processing peer to facilitate the estimation in the composite peer of the transmitted data signals from the processing peer. Said downlink combinatory processing methods produce a spatial diversity sample, on basis of at least two data signals, resulting in combined data signals. Said combined data signals then undergo ISP, and are afterwards transmitted by the spatial diversity means, resulting in transmitted data signals, the spectra of said transmitted data signals being at least partly overlapping. The transmitted data signals are then transmitted on the channel. Said combinatory processing can be described as a step for determining combined data signals in said transmitting terminal, said combined data signals being transformed versions of said data signals. Thereafter inverse subband processing of said combined data signals is performed, followed by transmitting with said spatial diversity means said inverse subband processed combined data signals, the spectra of said transmitted inverse subband processed combined data signals being at least partly overlapping.

[0071] In an embodiment of the invention, said methods for combinatory processing for downlink communication apply per-subband combinatory processing for downlink communication in at least one subband. In this embodiment, said downlink combinatory processing methods produce a combined data signal in one subband, on basis of the data signals in that specific subband. This can be described as determining of combined data signals in said transmitting terminal on a subband by subband basis.

[0072] In an embodiment of the invention, said downlink combinatory processing methods calculate said combined data symbol, called Y[s] by linearly combining the data signals, called X[s], with a precompensation matrix $E_{PRE}[s]$. In this embodiment, the Formula 6 holds:

$$X[s] = H^T[s] \,.\, \underbrace{E_{PRE}[s] \,.\, Y[s]}_{Y_{TR}[s]} + N''[s] \qquad \text{Formula 6}$$

[0073] In an embodiment, U = A and $E_{PRE}[s]$ is the inverse of the channel matrix $H^{T(-1)}[s]$.

[0074] In another embodiment, U < A and $E_{PRE}[s]$ is the pseudo-inverse of the channel matrix such that Formula 7 holds:

$$H^T[s].\,E_{PRE}[s] = I_{U \times U} \qquad \text{Formula 7}$$

[0075] In an embodiment of the invention, said methods for combinatory processing for downlink communication take into account non-idealities in the analogue front-end(s) in the transmitting- and or receiving terminal(s).

[0076] In an embodiment of the invention, said methods for combinatory processing for downlink communication avoid distortion in the analogue front-end(s).

[0077] In an embodiment of the invention, said precompensation matrix comprises nulls for one or more data signals on one or more specific subbands. In this embodiment, efficient and effective power usage can be obtained.

[0078] In an embodiment of the invention, as well the terminals in the processing peer as the terminals in the composite peer dispose of spatial diversity means and of subband and combinatory processing means, and both the processing and the composite peers embody at least two terminals transmitting data signals at least partially simultaneously whereby these transmitted data signals have at least partially overlapping spectra. In this embodiment, both the uplink and the downlink combinatory processing methods can be used in either transmission direction between two peers.

[0079] In a preferred embodiment, the processing peer is the basestation of a star-configuration network, which can be connected to the backbone of a wired network. Different terminals communicate simultaneously in the same frequency band with the basestation. The terminals only need a single front-end, and a plain OFDM-modem (i.e. without SDMA processing capabilities). This embodiment can work in a multipath fading environment, for applications such as for example Wireless Local Area Networks. This embodiment can work in the 5-6 GHz band, and can achieve a network capacity of 155 Mbps and above. In this embodiment, the spatial diversity means consisting of multiple transmitting and/or receiving means are concentrated in the basestation, thus reducing the overall hardware complexity and cost of the configuration. In this embodiment, the spatial diversity means comprises of multiple antennas spaced half a wavelength apart. Also, the peers communicate on basis of OFDM as a modulation technique. The ISP and SP are

performed on basis of IDFT (Inverse Discrete Fourier Transform) and DFT (Discrete Fourier Transform) processing respectively, and the subbands can be referred to as carriers. A guard interval containing a cyclic prefix is inserted between OFDM symbols at the transmitter side. SDMA is then applied in the frequency-domain. In addition to the aggregate of the advantages of both OFDM and SDMA, this approach results in simpler exploitation of spatial diversity compared to time-domain methods. For the estimation algorithms of the OFDM/SDMA systems, known least-squares (LS) or maximum likelihood symbol estimation (MLSE) algorithms can be applied. Also a novel class of uplink OFDM/SDMA methods with improved performance and still low complexity can be applied. These methods implement successive interference cancellation per carrier, state insertion and coherence grouping.

[0080] To illustrate the performance, the algorithms are applied to a 100 Mbps OFDM/SDMA WLAN. It has a 4-antenna basestation that separates up to 4 simultaneous users by SDMA. Each of these transmit 256-OFDM symbols at a datarate of 25 Mbps with QPSK. The guard interval is designed to comprise all the echoes received on the multipath propagation channel, so that the channel convolution becomes cyclic after removal of the guard interval. Thus, in the frequency-domain it becomes equivalent to multiplication with the Fourier transform of the channel, $h^U{}_a[s]$. Measures are taken to synchronize the different users communicating simultaneously in the same band. Under these conditions, the received data Y[s] can be written as in Formula 8. It is possible under the given circumstances to calculate an estimation $\bar{x}^u$ [s] of the data sequences $X^U[s]$ on a carrier per carrier basis. This per carrier estimation greatly simplifies the SDMA processing, and it allows a profound parallellization of this processing. Indeed an apparatus for performing SDMA/OFDM can comprise of a plurality of circuits being adapted for determining estimates of data signals based on part of the subbands of the subband processed received data signals, preferably one subband per circuit. Subsequently to the per carrier estimation, a decision on which element of the transmitting alphabet is nearest to the signal(s) resulting from the per carrier estimation. Said decision results in hard estimates $\tilde{x}^U$.

$$\underbrace{\begin{bmatrix} y_1[s] \\ \vdots \\ y_A[s] \end{bmatrix}}_{Y[s]} = \underbrace{\begin{bmatrix} h_1^1[s] \ldots h_1^U[s] \\ \vdots \qquad \vdots \\ h_A^1[s] \ldots h_A^U[s] \end{bmatrix}}_{H[s]} \cdot \underbrace{\begin{bmatrix} x^1[s] \\ \vdots \\ x^U[s] \end{bmatrix}}_{X[s]} + \underbrace{\begin{bmatrix} n_1[s] \\ \vdots \\ n_A[s] \end{bmatrix}}_{N[s]} \qquad \text{Formula 8}$$

[0081] The per carrier SDMA processing in the uplink can be performed in a linear fashion. In this case, the estimates of the data signals transmitted by the terminals $\tilde{x}^U$ [s] are calculated by linearly combining the single corresponding carrier signals received on the different antennas with the equalizer coefficients E[s], following Formula 1.

[0082] Said E[s] is calculated to minimize the expectations given in Formula 2. For a given noise energy (sigma squared) and conditions on x given by Formula 3, E[s] obeys the U sets of linear equations of Formula 4, wherein the superscript H denotes the Hermitian transpose.

[0083] Subsequently to equalization, the soft estimates are fed into a slicer, which determine the nearest constellation points with a decision method. This results in the hard estimates $\tilde{x}^U$. The equalizer coefficients E[s] could alternatively be calculated to maximally annihilate the channel distortion without taking the noise energy into account, in a so-called zero-forcing way. In this method, E[s] obeys the U sets of linear equations, given by Formula 5.

[0084] The performance of the system is evaluated by simulation with realistic channel data obtained from ray-tracing. The resulting curves show the bit error rate (BER) as a function of the received signal-to-noise ratio per bit (SNR).

[0085] Figure 3 shows the performance of LS-OFDM/SDMA for one to four simultaneous users. As a reference, the dashed curve gives the performance of a single-user single-antenna 100 Mbps plain OFDM link. An important observation is that the four-antenna four-user LS-OFDM/SDMA system outperforms plain OFDM. This demonstrates that a bandwidth re-use factor four is achievable without any performance penalty.

[0086] To evaluate the implementation complexity, we determine the total number of operations needed to execute the LS-OFDM/SDMA algorithm. The overall functionality can be separated into initialization and processing. During initialization, the equalizers $E_{LS}[s]$ are calculated from Formula 4 by Gaussian elimination with multiple right-hand sides. During processing, equalization and slicing have to be performed, which respectively correspond to a matrix multiplication and a set of comparators. Note that processing is done continuously, at the symbolrate, opposed to the initialization which is only calculated once. Figure 9 summarizes the approximative number of multiplications, additions and data transfers needed for the execution of both phases, per sub-carrier. The number of data transfers is an indicative number for the amount of memory/register transfers. Since the architecture has not been determined yet, the allocation of the data transfers to memory or register banks is an open issue. However, it is important to stress that data transfers often are the implementation bottleneck. As an example, the four-user four-antenna system from would require 72

kflops and 200kdata transfers during initialization and 400 Mflops/sec and a data transfer bandwidth of 1.2 Gwords/sec during processing.

[0087] In an OFDM/SDMA system, for certain subcarriers one or more users may be completely buried in multi-user interference (MUI) or have highly correlated channel vectors. Obviously, these users will suffer from residual MUI and noise after linear equalization. To mitigate this effect, in the invention Successive Interference Cancellation (SIC) is used, preferably Per Carrier (pcSIC). Opposed to LS-OFDM/SDMA, this technique does not estimate all users simultaneously, but does this successively, preferably on a certain subcarrier. As such the MUI originating from users that have been detected already, can be removed (thereby modifying the original signal carrier). This technique relies on feedback of intermediate hard estimates and is thus non-linear. Note also that pcSIC-OFDM/SDMA -since it works preferably on a per-carrier basis- elegantly exploits frequency diversity, which is another example of synergy from the OFDM/SDMA combination. Further per-carrier successive interference cancellation (pcSIC) is presented, although the invention is not limited hereto. On each subcarrier n, the detection order is first determined, preferably according to the received signal power. We assume that user 1 to user U are properly ordered. Successively, each user's soft estimate is calculated by least-squares linear combining according to Formula 9.

$$\tilde{x}^u[s] = \underbrace{\left[e_1^u[s] \cdots e_A^u[s]\right]}_{E_{LS}^u[s]} \cdot \begin{bmatrix} y_1^u[s] \\ \vdots \\ y_A^u[s] \end{bmatrix} \qquad \text{Formula 9}$$

$$H^u[s] = \begin{bmatrix} h_1^u[s] & \dots & h_1^U[s] \\ \vdots & & \vdots \\ h_A^u[s] & \dots & h_A^U[s] \end{bmatrix} \qquad \text{Formula 10}$$

[0088] In this equation $E_{LS}[s]$ is found as in Formula 4 except that for each iteration u, H[s] is replaced by Formula 10 with the $y^1{}_a[s]$ equal to the initial $y_a[s]$ from Formula 8. Afterwards, the MUI originating from user u is reconstructed (in general called recombining) and subtracted (thereby modifying) from the residual MUI as in Formula 11 with $\bar{x}^U[s]$ the hard intermediate estimate of user u after slicing (being a decision method). In the plain pcSIC algorithm, there is only one intermediate hard estimate for each data signal on each carrier, which will consequently be selected as the hard estimate (being a selection method).

$$\begin{bmatrix} y_1^{u+1}[s] \\ \vdots \\ y_A^{u+1}[s] \end{bmatrix} = \begin{bmatrix} y_1^u[s] \\ \vdots \\ y_A^u[s] \end{bmatrix} - \begin{bmatrix} h_1^u[s] \\ \vdots \\ h_A^u[s] \end{bmatrix} \bar{x}^u[s] \qquad \text{Formula 11}$$

[0089] Successive Interference Cancellation (SIC) can be described as an uplink transmission method wherein said determining of said estimates of said data signals from said subband processed received data signals in said receiving terminal for at least one data signal (but not limited to one) the following steps: (first step) selecting from said data signals a selected data signal. This selecting step can for instance be but is not limited to selecting the first data signal from the data signals being ordered according to received signal power. (second step) determining an estimate of said selected data signal from said subband processed received data signals, for instance but not limited to linear combining of the subband processed received data signals including slicing. (third step) modifying said subband processed received data signals based on said estimate of said selected data signal, for instance but not limited to recombining

and subtraction according to Formula 11, such that modified subband processed received data signals are obtained. (fourth step) finally estimates of said remaining data signals from said modified subband processed received data signals are determined, possibly by applying the same steps successively. Note that subband can also be denoted carriers.

[0090] In an embodiment of the successive interference cancellation uplink transmission method said determining of estimates (all steps) is performed on a subband by subband basis.

[0091] Figure 4 shows the performance of pcSIC-OFDM/SDMA. Again, the BER over SNR curves are given for one to four users. As a reference, the dashed curve gives the performance of single-user single-antenna 100 Mbps plain OFDM and the thin curves give the performance of LS-OFDM/SDMA. It shows that pcSIC-OFDM/SDMA yields a performance improvement compared to LS-OFDM/SDMA that increases with an increasing number of users. To be specific, for a BER of 0.001 and for four simultaneous users, we observe a 5 dB gain. For a complexity estimation of the pcSIC algorithm for OFDM/SDMA, we also discern between initialization and processing. During initialization, the cancellation order is determined and the equalizers are calculated. Since the channel matrix is different for each user u the latter would require U distinct Gaussian eliminations. However, by exploiting structural relationships between the $H^u[s]$'s, we managed to reduce the required number of operations with $O(A^2\ U^2)$. During processing, reconstruction (recombination) and subtraction (modification) have to be performed in addition to equalization and slicing. The total approximative number of operations for both phases of the pcSIC-OFDM/SDMA algorithm are given in Figure 10 . More specific, the four-user four-antenna system would require respectively 170kflops and 250kdata-transfers in the initialization phase and 700Mflops/sec and a data-transfer bandwidth of 1.8Gwords/sec.

[0092] A potential weakness of (pc)SIC-OFDM/SDMA is that its performance degrades when at least two users are received with approximately equal power. Under such transmission conditions, the probability of making an erroneous decision is increased, resulting in error propagation. This potential deficiency is illustrated in Figure 5 which shows for each carrier -in the bottom part-the signal-to-interference ratio (SIR) during the first iteration and -in the upmost part-the number of errors that occurred. Note however that from an information-theoretic viewpoint it is optimal that all users have identical power. To resolve the problem of error propagation, we use interference-dependent state insertion (SI). Essentially, by inserting additional state information on those carriers that suffer from a bad SIR, we can seriously decrease the probability of error propagation at a reasonable cost. Preferably State Insertion is applied in the (pc)SIC-OFDM/SDMA methods. The State Insertion method is further presented in a subcarrier by subcarrier approach but is not limited hereto.

[0093] First, the SIRs of user 1 for each subcarrier n are calculated from the knowledge of the equalizers $E_{LS}[s]$. Next, one additional state m is assigned to each of the M subcarriers n_m that suffer most from SIR. These M states keep track of alternative estimates, called intermediate hard estimates in the above, $\bar{x}_M^1[m]$ for user 1. These are defined as the nearest constellation point to $\tilde{x}_M^1[s_m]$ except for $\bar{x}^1[s_m]$, which is the sliced version of $\tilde{x}^1[s_m]$. After these assignments the M additional states are treated as normal subcarriers to obtain $\bar{x}_M^u[s_m]$, successively for u=2 to U. In explicit, the MUI originating from user u-1 is reconstructed (after recombining with a channel response estimation) and subtracted (thereby modifying the original signal) according to Formula 12 and the soft estimates $\tilde{x}_M^u[s_m]$ are computed by least-squares combining according to Formula 13. Finally, for each of those M subbands the algorithm selects from the intermediate hard estimates either $\bar{x}^u[s_m]$ or $\bar{x}_M^u[s_m]$ as hard estimate if respectively the first or second of the 2-norms of Formula 13 is smallest.

$$\begin{bmatrix} y_{1_M}^u[m] \\ \vdots \\ y_{A_M}^u[m] \end{bmatrix} = \begin{bmatrix} y_{1_M}^{u-1}[m] \\ \vdots \\ y_{A_M}^{u-1}[m] \end{bmatrix} - \begin{bmatrix} h_{1_M}^{u-1}[s_m] \\ \vdots \\ h_{A_M}^{u-1}[s_m] \end{bmatrix} \bar{x}_M^{u-1}[m] \qquad \text{Formula 12}$$

$$\tilde{x}_M^u[m] = \left[e_1^u[s_m] \cdots e_A^u[s_m]\right] \cdot \begin{bmatrix} y_{1,M}^u[m] \\ \vdots \\ y_{A,M}^u[m] \end{bmatrix} \qquad \text{Formula 13}$$

[0094] The State Insertion uplink transmission method can be described as a method wherein said determining of said estimates of said data signals from said subband processed received data signals in said receiving terminal further comprising for at least one data signal (but not limited to one) the following steps: (first step) selecting from said data signals a selected data signal; (second step) determining a plurality of estimates of said selected data signal from said subband processed received data signals, said estimates are above denoted intermediate hard estimates; (third step) determining a plurality of modified subband processed received data signals, each of said modified subband processed received data signals being based on one of said (intermediate hard) estimates of said selected data signal. A recombination and modifier method are exploited here; (fourth step) determining a plurality of estimates of at least one of said remaining data signals from each of said plurality of modified subband processed received data signals, for instance via but not limited to least squares linear combination; (last step) and thereafter selecting one of said estimates of said selected data signal, for instance but not limited to evaluation of a norm of said modified subband processed received data signals.

[0095] In an embodiment of State Insertion uplink transmission method said determining of estimates (all steps) is performed on a subband by subband basis.

[0096] In an embodiment of the invention said State Insertion uplink transmission method is used in combination with Successive Interference Cancellation, preferably on a subband by subband basis.

[0097] Figure 6 shows the performance of pcSIC-OFDM/SDMA with interference-dependent state insertion for the exemplary system. The number of extra states was set to M=64. As a reference, the dashed curve gives the performance of single-user single-antenna 100Mbps plain OFDM and the thin curves give the performance of pcSIC-OFDM/SDMA without state insertion. It shows that pcSIC-OFDM/SDMA with 64 additional states achieves a 6 dB gain at a BER of 0.001 and for four simultaneous users compared to pcSIC-OFDM/SDMA without SI. Other simulations also show that the performance improvement for SI with 32 and 16 states is respectively 5 dB and 4 dB.

[0098] For an estimation of the complexity of pcSIC-OFDM/SDMA with SI, the functionality can again be split in an initialization and a processing part. During initialization, each subcarrier's SIR after equalization needs to be computed, followed by state insertion for the M worst SIR subcarriers. During processing, M extra states need to be tracked using Formula 12 and 13, followed by selection based on the residual 2-norms of Formula 14.

$$\left\| \begin{matrix} y_1^{U+1}[s_m] \\ \vdots \\ y_A^{U+1}[s_m] \end{matrix} \right\|_2 \quad or \quad \left\| \begin{matrix} y_{1_M}^{U+1}[m] \\ \vdots \\ y_{A_M}^{U+1}[m] \end{matrix} \right\|_2 \qquad \text{Formula 14}$$

[0099] The operations required for state insertion of M out of N subcarriers are given in Figure 11. To obtain the total number of operations for pcSIC-OFDM/SDMA with SI the operations from Figure 10 have to be added to this. For the four-user four-antenna system, pcSIC-OFDM/SDMA with 64 additional states would require 220kflops and 310kdata-transfers during the initialization phase and 1.1Gflops/sec and a data-transfer bandwidth of 2.6Gwords/sec in the processing phase.

[0100] In the downlink transmission, the terminals of the composite peer are receiving signals and the processing peer is transmitting signals. The receiving terminals only dispose of a single antenna and a plain OFDM modem (i.e. without SDMA processing capabilities) in this embodiment. Since the user terminals have only a single antenna and since we want to concentrate most of the processing power in the basestation, they cannot mitigate multi-user interference in the downlink. Therefore, the basestation carries out a precompensation matrix E_{PRE} on the datasymbol vector Y[s] to obtain $Y_{TR}[s]$, which is then transmitted on the A antennas. The input-output relation of Formula 6 is obtained.

[0101] Neglecting any non-idealities, this approach results in perfect separation of each user's datasymbols and perfect equalization upon reception. Therefore, no channel information or equalizer is needed in the mobiles. By applying these precompensation matrices for each carrier n, the input-output relation becomes as in Formula 15.

$$\begin{aligned} X[s] &= H^T[s] \cdot E_{PRE}[s] \cdot Y[s] + N''[s] \\ &= I_{U \times U} \cdot Y[s] + N''[s] \end{aligned} \qquad \text{Formula 15}$$

[0102] Thus, each user sees a single-user AWGN channel per carrier. Figure 7 shows the average BER in the downlink for different numbers of users. Since there is no interference, the number of simultaneous users does not influence performance (provided it does not exceed the number of antennas and no non-idealities occur).

[0103] The processing in the downlink again consists of both initialization and processing. In the initialization phase, the pre-compensation matrices are calculated. During the processing, the user data Y[s] is then multiplied with said pre-compensation matrices. The complexity for both initialization and processing is approximately the same as for the LS-OFDM/SDMA scheme in the uplink.

[0104] In addition to the proposed OFDM/SDMA techniques proposed for both uplink and downlink communication, the initialization effort in both situations can be simplified by applying coherence grouping. This initialization effort is characterized as determining the equalizer or the precompensation matrices, wherein Formulas 4, 5 or 7 are exploited. With said equalizer matrices relations between said data signals and subband processed received data signals are defined like in Formula 1, 9 or 13. With said precompensation matrices relations between said combined data signals, being transformed versions of said data signals, are defined like in Formula 6 or 15. Instead of determining said matrices for each subband separately, said subbands can be grouped into sets, at least one set comprising of at least two subbands and said matrices or more generally said relations can then be determined on a set-by-set basis. The groups are in an embodiment of this coherence grouping principle all of equal size G. Figure 8 shows the performance degradation associated with carrier grouping applied to pcSIC-OFDM/SDMA with SI, for several values of G. It is observed that for the simulated channel the degradation is negligible for G smaller then or equal to 4. For G greater then or equal to 32, the performance becomes inferior to single-user single-antenna plain OFDM. We may conclude that for the system of interest the initialization complexity can be reduced with a factor 8. For the example of a 100 Mbps OFDM/SDMA WLAN that uses SDMA to separate four simultaneous 25 Mbps users, some numerical results for the performance gain and the complexity can be given. If pcSIC-OFDM/SDMA with 64-SI and a group size of eight is implemented, a performance gain of 11dB is obtained at a BER of 0.001 compared to the least-squares approach. The required computational power is respectively 27kflops and 40kdata-transfers during initialization and 1.1Gflops/sec and a data-transfer bandwidth of 2.6Gwords/sec during processing.

[0105] In another preferred embodiment, the processing peer provides an access point to a connecting network (this connecting network can for example be a cable network or a satellite network), and the composite peer consists of wireless terminals for which cost is a major issue. In this embodiment, only the access point has DFT (Discrete Fourier Transform) and IDFT (Inverse Discrete Fourier Transform) processing means. This enables a cheap realization of both the digital baseband modem part of the terminals in the composite peer, and of the front-ends of all terminals in the system since a high peak-to-average power ratio of the transmitted signals is avoided. A typical application of asymmetric frequency domain SDMA could for example be a wireless home networking application in the 2.4 GHz band. Such an asymmetric configuration does implements a so-called concentrated scenario, wherein subband processing and inverse subband processing is located in the processing peer. In this transmission method the determination of estimates of said data signals from said subband processed received data signals in said receiving terminal comprising the following steps: (first step) determining intermediate estimates of said data signals from said subband processed received data signals in said receiving terminal; (second step) obtaining said estimates of said data signals by inverse subband processing said intermediate estimates. In this embodiment, the methods of the invention can be used not only to detect the signals of different terminals transmitting simultaneously in the same frequency band, but for example also to mitigate the interference from a microwave oven causing disturbance signals in the band. In multipath fading environment, such as for example home environments, guard intervals can be inserted in the data signals prior to transmission. If these guard intervals are again designed to comprise all the echoes received on the multipath channel, the channel convolution becomes cyclic after removal of the guard intervals. Thus, in the frequency-domain it becomes equivalent to multiplication with the Fourier transform of the channel. This embodiment allows cheap terminals in the processing peer and can implement uplink as well as downlink combinatory processing, in a way similar as OFDM/SDMA, including successive interference cancellation, preferably per subband and state insertion.

Claims

1. A method of transmitting data signals (50) from at least two transmitting terminals (20) with each at least one transmitting means (60) to at least one receiving terminal (40) with a spatial diversity receiving means (80) comprising the steps:

    - transmitting from said transmitting terminals (20) transformed data signals (70), being transformed versions of said data signals; receiving on said spatial diversity means (80) received data signals being at least function of at least two of said transformed data signals (70);
    - subband processing (90) of at least two of said received data signals in said receiving terminal (40); and

- determining estimates of said data signals (120) from subband processed received data signals (140) in said receiving terminal.

2. The method recited in 1, wherein said transmissing being substantially simultaneously.

3. The method recited in 1, wherein, the spectra of said transformed data signals being at least partly overlapping.

4. The method recited in 1, wherein the spectra of said transformed data signals being at least partly overlapping.

5. The method recited in 1, wherein the step of determining of said estimates of said data signals from subband processed received data signals in said receiving terminal further comprising for at least one data signal the steps of:

   - selecting from said data signals a selected data signal;
   - determining an estimate of said selected data signal from said subband processed received data signals;
   - modifying said subband processed received data signals based on said estimate of said selected data signal; and
   - determining estimates of said remaining data signals from said modified subband processed received data signals.

6. The method recited in claim 5, wherein the step of selecting a data signal being based on the receiving power of said data signals.

7. The method recited in claim 5, wherein the step of selecting a data signal being based on the interference ratio of said data signals.

8. The method recited in claim 1, wherein the step of determining of said estimates of said data signals from subband processed received data signals in said receiving terminal further comprises for at least one data signal the steps of:

   - selecting from said data signals a selected data signal;
   - determining a plurality of estimates of said selected data signal from said subband processed received data signals;
   - determining a plurality of modified subband processed received data signals, each of said modified subband processed received data signals being based on one of said estimates of said selected data signal;
   - determining a plurality of estimates of at least one of said remaining data signals from said plurality of modified subband processed received data signals; and
   - thereafter selecting one of said estimates of said selected data signal.

9. The method recited in claim 8, wherein the step of selecting a data signal being based on the interference ratio of said data signals.

10. The method recited in claim 1, wherein:

   - the subbands, being involved in said subband processing, being grouped into sets, at least one set comprising of at least two subbands;
   - the step of determining of said estimates of said data signals in said receiving terminal comprising the steps:
   - determining relations between said data signals and subband processed received data signals on a set-by-set basis;
   - exploiting said relations between said data signals and said subband processed received data signals for determining said data signals.

11. The method recited in claim 1, wherein said transformation of said data signals (50) to transformed data signals (70) comprising the step of inverse subband processing (160).

12. The method recited in claim 1, wherein the step of determining (150) estimates of said data signals from subband processed received data signals in said receiving terminal comprises the steps of:

   - determining (100) intermediate estimates of said data signals (130) from said subband processed received

data signals in said receiving terminal;

- obtaining said estimates of said data signals (120) by inverse subband processing (110) said intermediate estimates.

13. The method recited in claim 1, wherein said transformation of said data signals to transmitted data signals further comprising guard interval introduction.

14. The method recited in claim 1, wherein said subband processing being orthogonal frequency division demultiplexing.

15. The method recited in claim 11 and 12, wherein said inverse subband processing being orthogonal frequency division multiplexing.

16. A method of transmitting data signals (200) from at least one transmitting terminal (240) with a spatial diversity transmitting means (220) to at least two receiving terminals (330)with at least one receiving means (320) comprising the steps:

- determining (250) combined data signals (300) in said transmitting terminal, said combined data signals being transformed versions of said data signals;
- inverse subband processing (260) said combined data signals;
- transmitting with said spatial diversity means (220) inverse subband processed combined data signals;
- receiving on at least one of said receiving means (320) of at least one of said receiving terminals (330) inverse subband processed received data signals, being at least function of said inverse subband processed combined data signals; and
- determining estimates of said data signals from said inverse subband processed received data signals.

17. The method recited in claim 16, wherein said transmitting of inverse subband processed combined data signals being substantially simultaneously.

18. The method recited in claim 16, wherein the spectra of said inverse subband processed combined data signals being at least partly overlapping.

19. The method recited in claim 16, wherein the step of determining of combined data signals in said transmitting terminal being on a subband by subband basis.

20. The method recited in claim 16, wherein the step of determining of said estimates of said data signals in said receiving terminals comprising subband processsing (350).

21. The method recited in claim 16, wherein the step of determining combined data signals in said transmitting terminal comprising the steps:

- determining intermediate combined data signals (290) by subband processing (280) said data signals;
- determining (270) said combined data signals from said intermediate combined data signals.

22. The method recited in claim 20 and 21, wherein said subband processing being orthogonal frequency division demultiplexing.

23. The method recited in claim 16, wherein said inverse subband processing being orthogonal frequency division multiplexing.

24. The method recited in claim 16, wherein:

- said subbands, being involved in inverse subband processing, being grouped into sets, at least one set comprising of at least two subbands;
- the step of determining (250) of combined data signals (300) in said transmitting terminal (240) comprising the steps:
- determining relations between said data signals and said combined data signals on a set-by-set basis; and
- exploiting said relations between said data signals and said combined data signals for determining said data

signals.

25. The method recited in claim 16, wherein in said inverse subband processed combined data signals a guard interval being introduced.

26. An apparatus for determining estimates of data signals from at least two received data signals, said received data signals, said apparatus comprising at least of

- at least one spatial diversity receiving means (80);
- circuitry being adapted for receiving said received data signals with said spatial diversity receiving means;
- circuitry being adapted for subband processing (90) at least two of said received data signals; and
- circuitry (150) being adapted for determining estimates of said data signals from subband processed received data signals.

27. The apparatus recited in claim 26 wherein said circuitry being adapted for determining estimates of said data signals from subband processed received data signals comprises a plurality of circuits each being adapted for determining part of said estimates of said data signals based on part of the subbands of said subband processed received data signals.

28. The apparatus recited in claim 26, wherein said spatial diversity means comprises of at least two receiving means and said circuitry being adapted for receiving said received data signals with said spatial diversity means comprises a plurality of circuits each being adapted for receiving said received data signals from one of said receiving means of said spatial diversity means.

29. An apparatus for transmitting inverse subband processed combined data signals comprising at least of:

- at least one spatial diversity transmitting means;
- circuitry being adapted for combining data signals;
- circuitry being adapted for inverse subband processing combined data signals;
- circuitry being adapted for transmitting inverse subband processed combined data signals with said spatial diversity means.

30. The apparatus recited in claim 29, wherein said circuitry being adapted for combining data signals comprising a plurality of circuits each being adapted for combining data signals based on part of the subbands of said data signals.

31. The apparatus recited in claim 29, wherein said spatial diversity transmitting means comprises at least two transmitting means and said circuitry being adapted for transmitting inverse subband processed combined data signals comprises a plurality of circuits each being adapted for transmitting said inverse subband processed combined data signals with one of said transmitting means of said spatial diversity means.

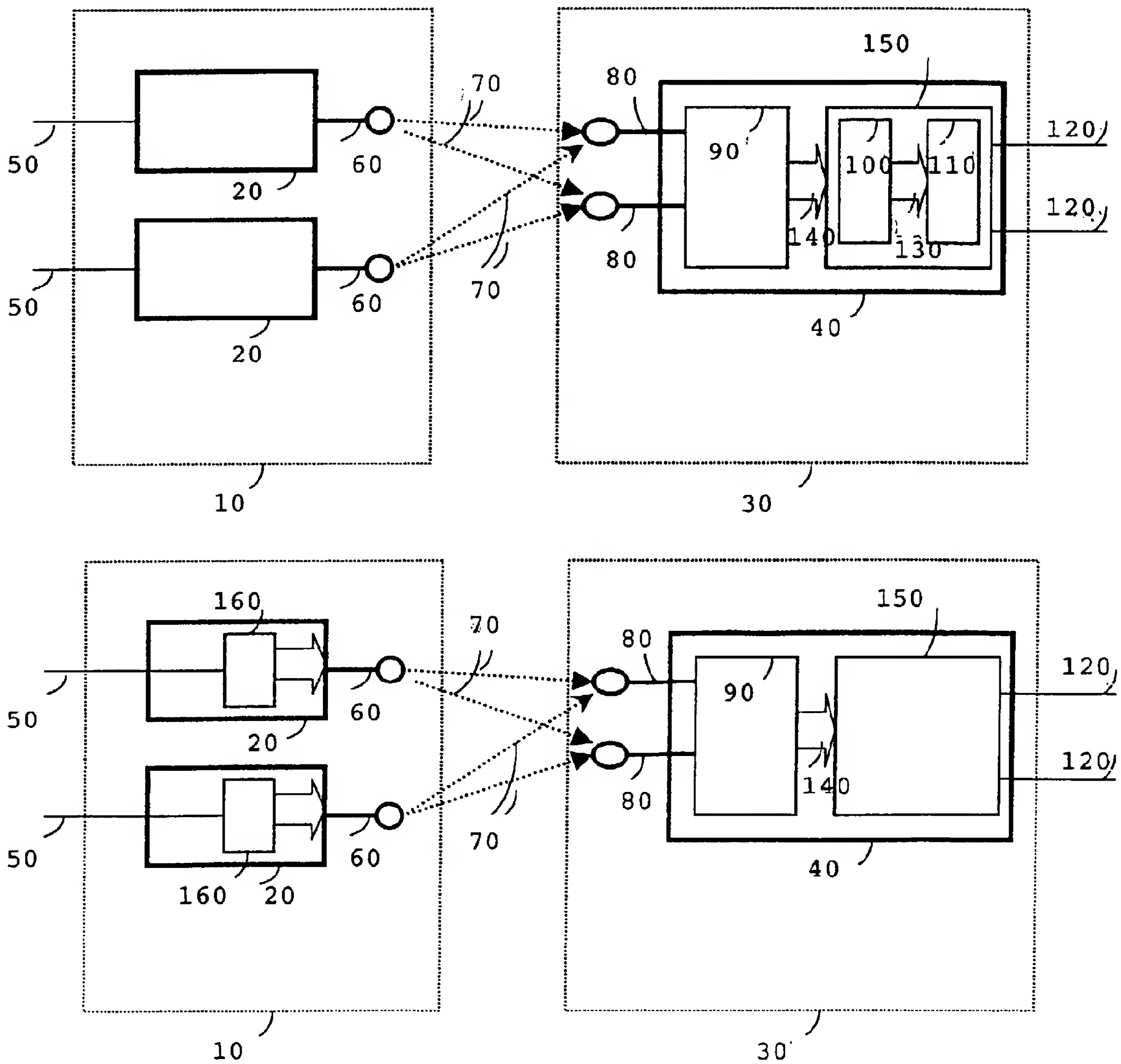
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FIG. 1

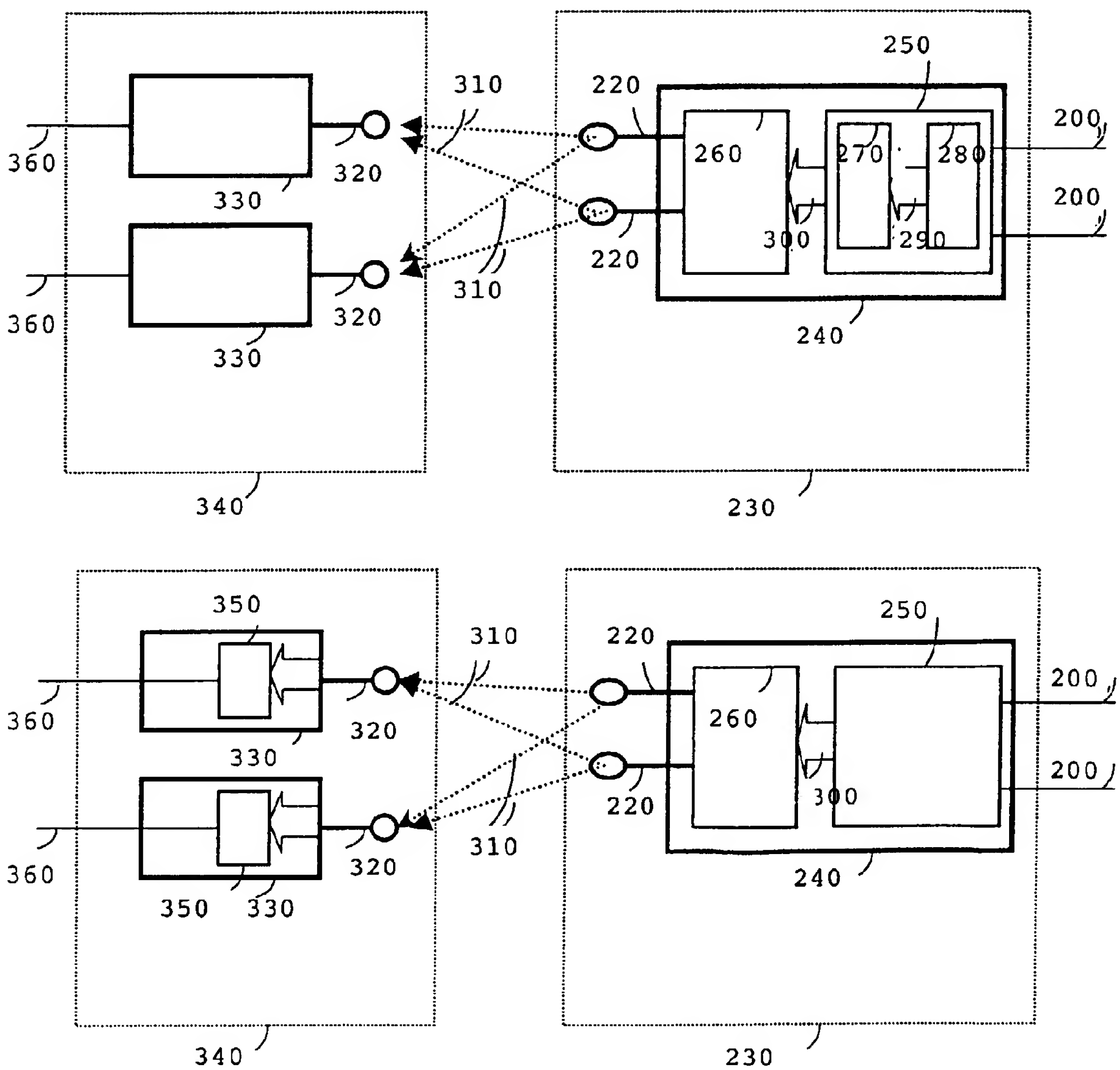
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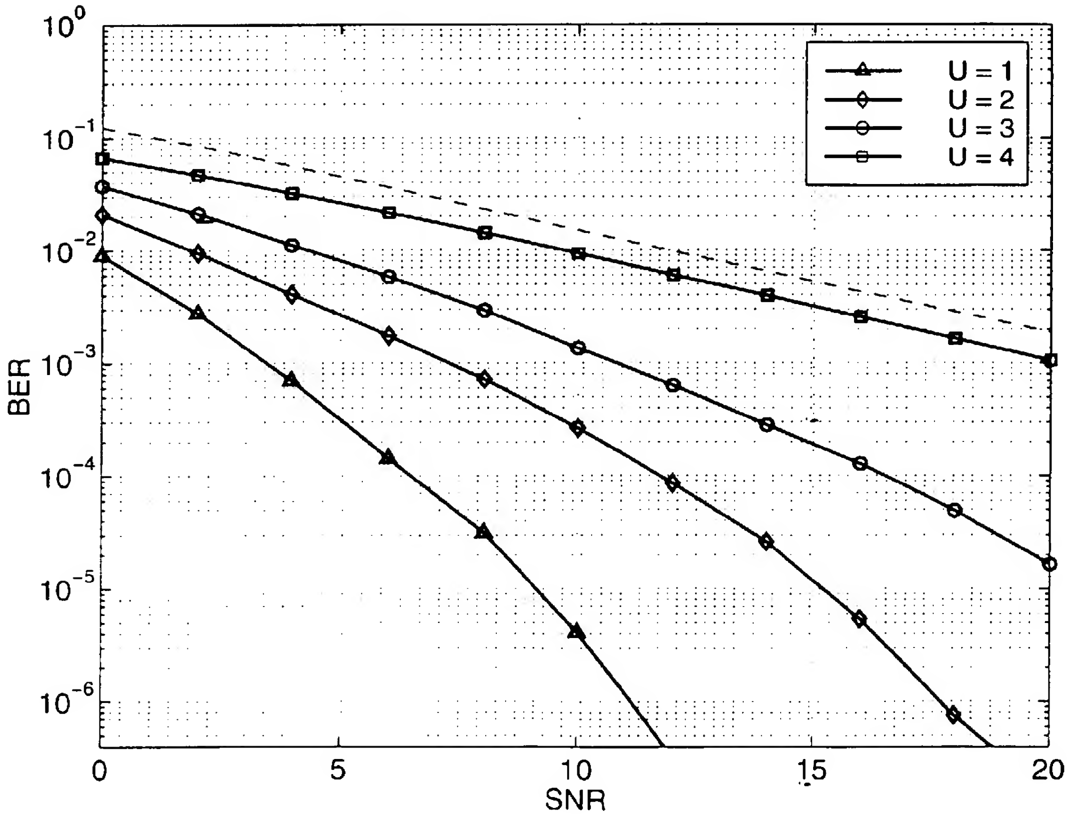
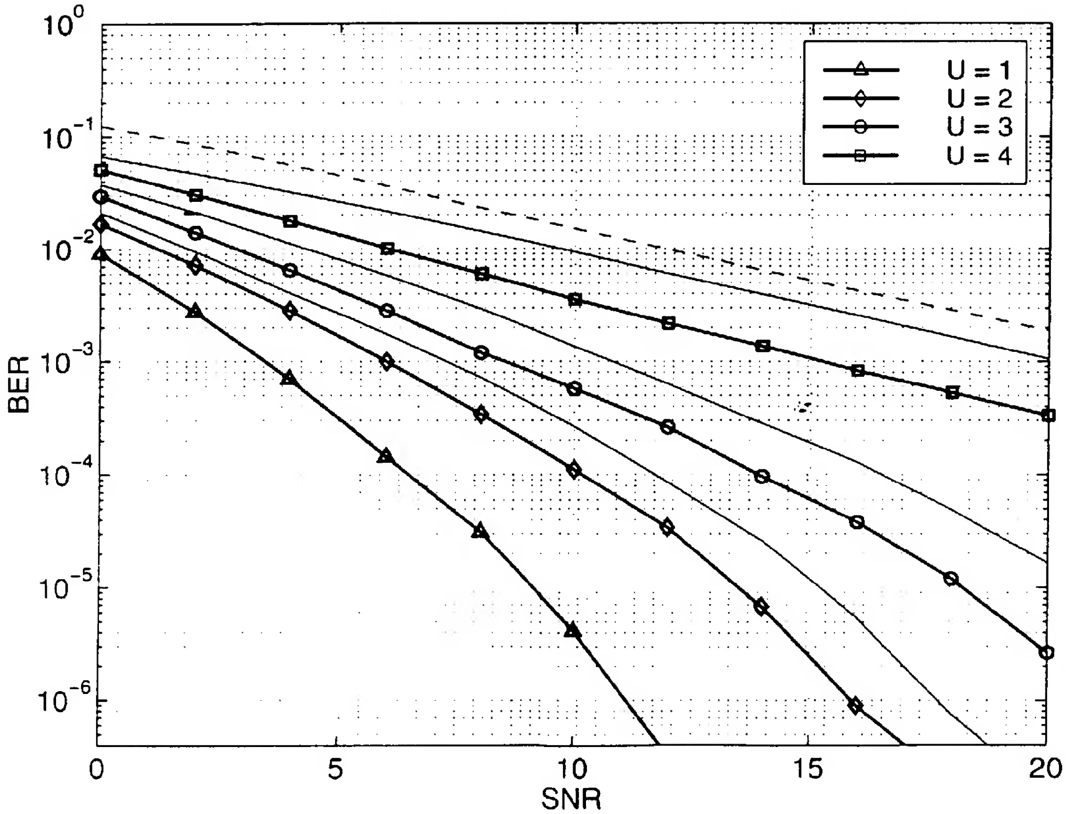
FIG. 2



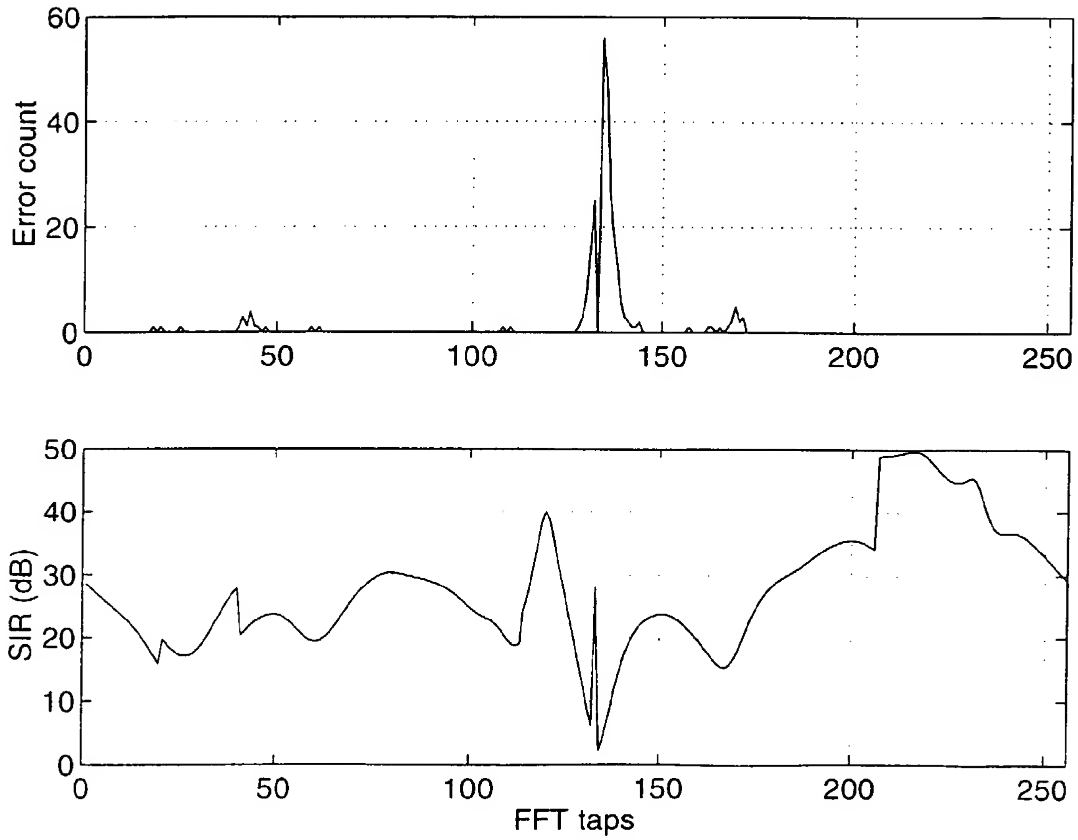
Performance of LS-OFDM/SDMA

FIG. 3



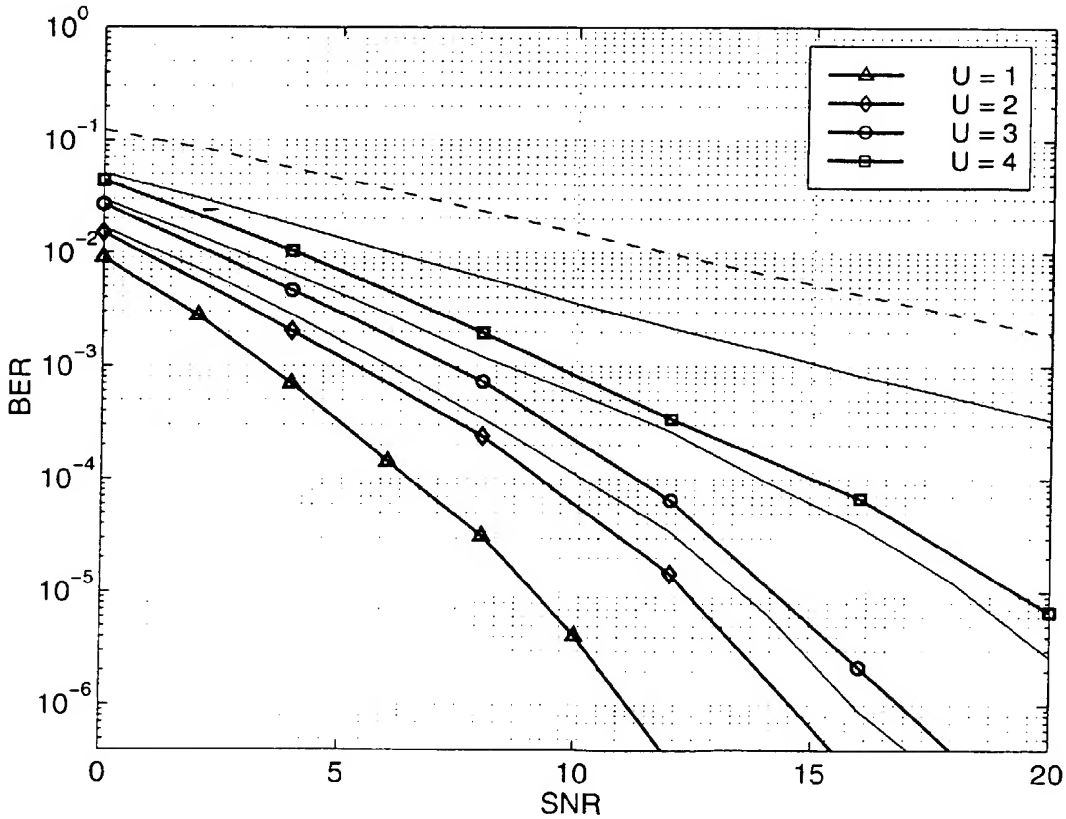
Performance of pcSIC for OFDM/SDMA

FIG. 4



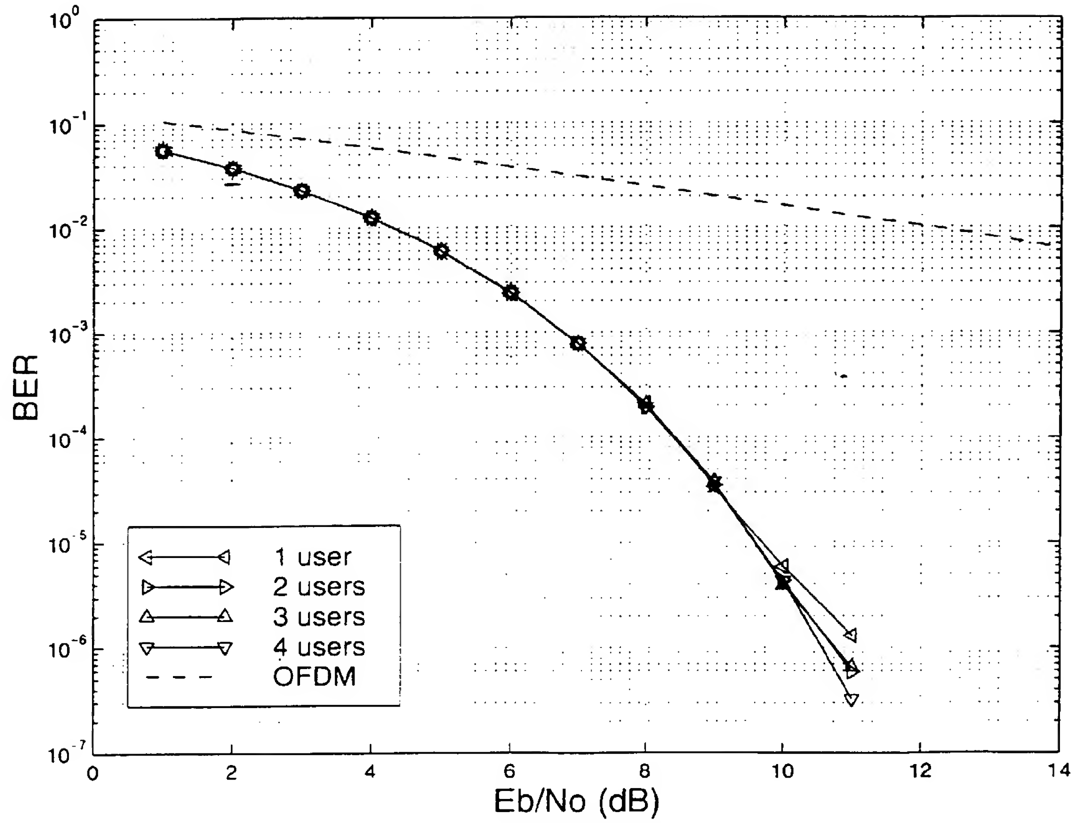
Error propagation with SIC-OFDM/SDMA

FIG. 5



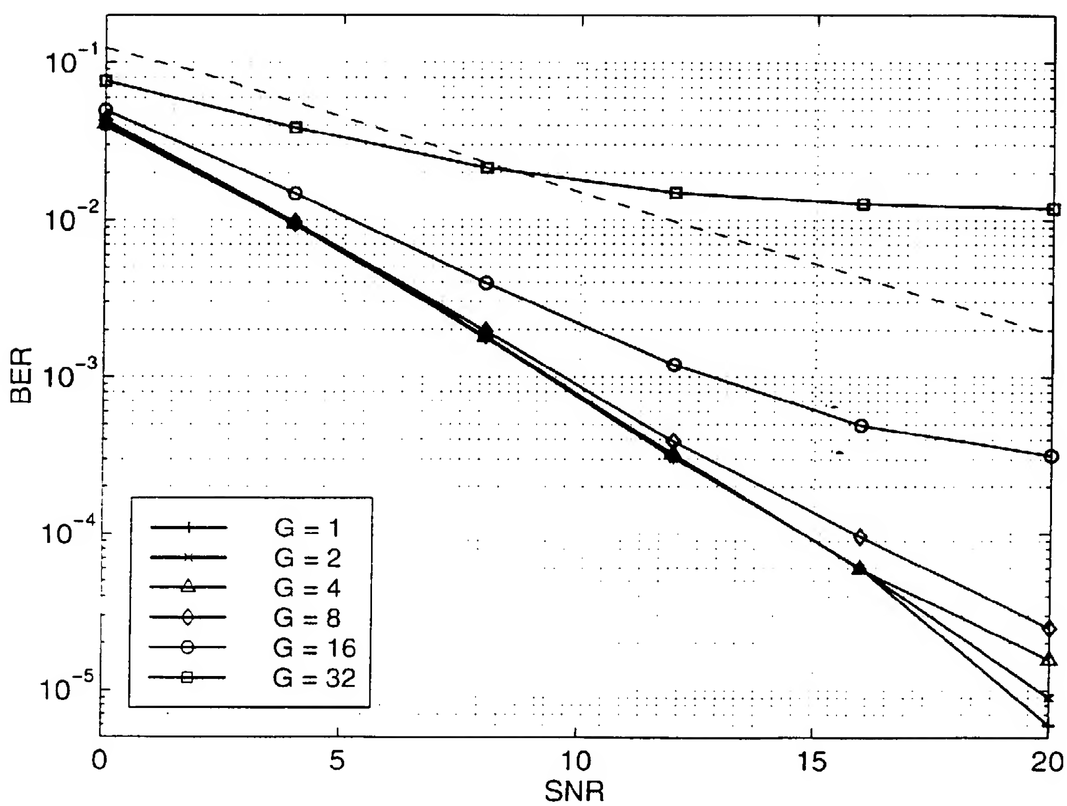
Performance of pcSIC-OFDM/SDMA with SI

FIG. 6



Performance of OFDM/SDMA in the downlink

FIG. 7



Coherency Grouping

FIG. 8

LS-OFDM/SDMA operation count

	initialization	processing
multiplications	$A^3/3 + 2A^2U$	AU
additions	$A^3/3 + 2A^2U$	AU
data transfers	$A^3 + 9A^2U$	$6AU$

FIG. 9

pcSIC-OFDM/SDMA operation count

	initialization	processing
multiplications	$A^3U/3 + 12A^2U$	$2AU - A$
additions	$A^3U/3 + 12A^2U$	$2AU - A$
data transfers	$A^3U/3 + 24A^2U$	$10AU - 4A$

FIG. 10

additional operation count for SI

	initialization	processing
multiplications	$AU^2/2$	$2(M/N\,AU + A)$
additions	$(AU^2 + N)/2$	$2(M/N\,AU + A)$
data transfers	$AU^2 + N/2$	$10(M/N\,AU + A)$

FIG. 11

European Patent Office

EUROPEAN SEARCH REPORT

Application Number
EP 99 87 0168

DOCUMENTS CONSIDERED TO BE RELEVANT

Category	Citation of document with indication, where appropriate, of relevant passages	Relevant to claim	CLASSIFICATION OF THE APPLICATION (Int.Cl.7)
X Y	WO 97 41647 A (BAIER PAUL WALTER ;KAMMERLANDER KARL (DE); JUNG PETER (DE); SIEMEN) 6 November 1997 (1997-11-06) * abstract * * page 2, line 27 - page 7, line 22 * * figure 1 *	1-3,11, 13-15,26 5,8,10, 12,27,28	H04B7/08 H04L27/26 H04L1/06
Y	FAZEL K. ; FETTWEIS G.P.: "Multi-Carrier Spread Spectrum" 1997 , KLUWER ACADEMIC PUBLISHERS , DORDRECHT, THE NETHERLANDS XP002130135 * page 50, paragraph 2 - page 52, paragraph 2 * * figures 1,2 *	16-25, 29-31	
Y Y	KUZMINSKIY, A.M.; HATZINAKOS, D.: "Multistage semi-blind spatio-temporal processing for short burst multiuser SDMA systems" CONFERENCE ON SIGNALS, SYSTEMS & COMPUTERS, vol. 2, 1 - 4 November 1998, pages 1887-1891, XP002130134 * abstract * * page 1888, column 1, paragraph 1 - page 1890, column 1, paragraph 1 *	5,8,10, 12,27,28 16-25, 29-31	TECHNICAL FIELDS SEARCHED (Int.Cl.7) H04B H04L
A	WO 98 51048 A (NEILL RORIE JULIAN TURNBULL O ;MOTOROLA LTD (GB)) 12 November 1998 (1998-11-12) * page 3, line 14 - page 4, line 27 * * page 7, line 14 - page 8, line 30 * * page 10, line 1-21 * * figures 3-5 *	1-3,11, 13-15,26	

The present search report has been drawn up for all claims

Place of search	Date of completion of the search	Examiner
THE HAGUE	9 February 2000	Yang, Y

CATEGORY OF CITED DOCUMENTS

X : particularly relevant if taken alone
Y : particularly relevant if combined with another document of the same category
A : technological background
O : non-written disclosure
P : intermediate document

T : theory or principle underlying the invention
E : earlier patent document, but published on, or after the filing date
D : document cited in the application
L : document cited for other reasons

& : member of the same patent family, corresponding document

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ANNEX TO THE EUROPEAN SEARCH REPORT ON EUROPEAN PATENT APPLICATION NO. EP 99 87 0168

This annex lists the patent family members relating to the patent documents cited in the above-mentioned European search report.
The members are as contained in the European Patent Office EDP file on
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09-02-2000

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		EP 0895683 A	10-02-1999
WO 9851048 A	12-11-1998	GB 2324932 A	04-11-1998
		AU 7648998 A	27-11-1998
		EP 0922350 A	16-06-1999

EPO FORM P0459

For more details about this annex : see Official Journal of the European Patent Office, No. 12/82

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(54) Transmit power control method and apparatus in a radio communication system

(57) When the SIR of the received signal is larger than the SIR upper limit, threshold determining section 105 outputs an instruction for decreasing the transmit power to control bit generating section 106. Further, when the SIR is smaller than the SIR lower limit, threshold determining section 105 outputs an instruction for increasing the transmit power to control bit generating section 106. When the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, threshold determining section 105 outputs the determined result to control bit generating section 106. When the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, control bit generating section 106 refers to a last instruction on the transmit power control stored in storage section 107, and outputs an instruction opposite to the last instruction with respect to an increase or decrease in the transmit power to storage section 107 and multiplexing section 108.

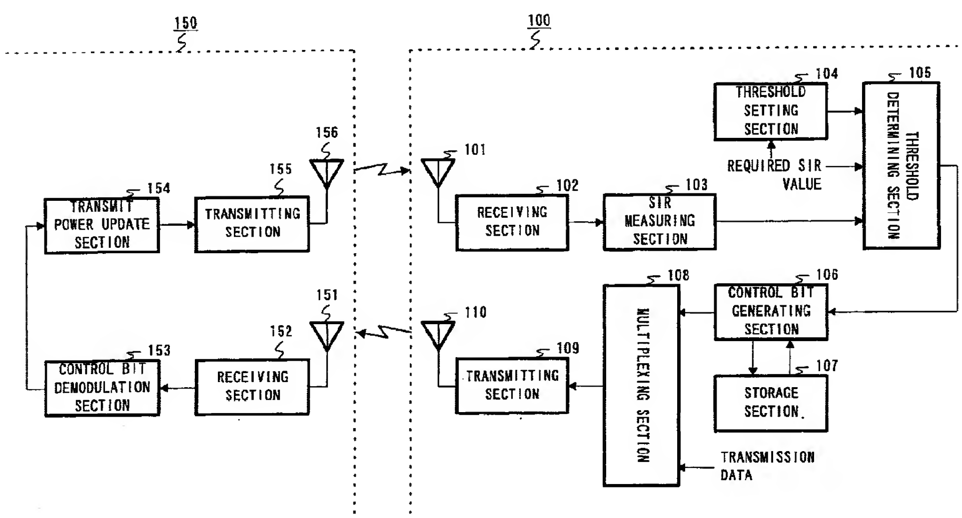

FIG. 3

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Description

BACKGROUND OF THE INVENTION

Field of the Invention

[0001] The present invention relates to a communication apparatus and transmit power control method, and more particularly, to a communication apparatus and transmit power control method used in a radio communication using a CDMA system as a radio access system.

Description of the Related Art

[0002] Recently, in a CDMA radio communication system, a case sometimes occurs that a plurality of control channels and user channels exists together in the same carrier frequency. Therefore, the existence of a channel on which signals are transmitted with the transmit power more than a required level causes the interference with other users, and thereby degrades the reception performance in the other users. Accordingly, it is necessary to suppress the transmit power to an extent enabling a predetermined received quality to be maintained.

[0003] A conventional transmit power control is herein explained. FIG.1 is a diagram illustrating a communication aspect of a base station apparatus and communication terminal apparatus.

[0004] Communication terminal apparatus 10 measures SIR (Signal to Interference Ratio) of a signal transmitted from base station apparatus 20. Then, communication terminal apparatus 10 compares the measured SIR with a required SIR value, and instructs base station apparatus 20 to increase or decrease the transmit power.

[0005] According to the instruction transmitted from communication terminal apparatus 10, base station apparatus 20 changes the transmit power of a signal to be transmitted to communication terminal apparatus 10.

[0006] However, as described in a variable rate CDMA transmit power control method disclosed in Japanese Laid Open Patent Publication HEI11-17646, in a conventional configuration, a transmit power control error sometimes occurs due to a time delay between the time the transmit power control information is determined and the time the transmit power is actually updated according to the transmit power control information.

[0007] An example of the above transmit power control will be described below. FIG.2 is a view showing an example of the conventional transmit power control.

[0008] In FIG.2, the abscissa is indicative of time, the ordinate is indicative of SIR, and instructions for communication terminal apparatus 10 to transmit to base station apparatus 20 are indicated at a lower portion in the figure. In FIG.2, the instruction "UP" is an instruction for increasing the transmit power of base station apparatus 20, while the instruction "DW" is another instruction for decreasing the transmit power of base station apparatus 20.

[0009] S1 is indicative of SIR of a signal received by communication terminal apparatus 10 at time t1. Since S1 has a value lower than a required SIR value a1, communication terminal apparatus 10 transmits the instruction "UP" to base station apparatus 20.

[0010] S2 is indicative of SIR of a signal received by communication terminal apparatus 10 at time t2. Since S2 has a value lower than the required SIR value a1, communication terminal apparatus 10 transmits the instruction "UP" to base station apparatus 20. Base station apparatus 20 receives at time t2 the instruction "UP" transmitted from communication terminal apparatus 10 at time t1, and increases the transmit power to transmit a signal to communication terminal apparatus 10.

[0011] S3 is indicative of SIR of a signal received by communication terminal apparatus 10 at time t3. Since S3 has a value lower than the required SIR value a1, communication terminal apparatus 10 transmits the instruction "UP" to base station apparatus 20. Base station apparatus 20 receives at time t3 the instruction "UP" transmitted from communication terminal apparatus 10 at time t2, and increases the transmit power to transmit a signal to communication terminal apparatus 10.

[0012] S4 is indicative of SIR of a signal received by communication terminal apparatus 10 at time t4. Since S4 has a value lower than the required SIR value a1, communication terminal apparatus 10 transmits the instruction "UP" to base station apparatus 20. Base station apparatus 20 receives at time t4 the instruction "UP" transmitted from communication terminal apparatus 10 at time t3, and increases the transmit power to transmit a signal to communication terminal apparatus 10.

[0013] S5 is indicative of SIR of a signal received by communication terminal apparatus 10 at time t5. Since S5 has a value higher than the required SIR value a1, communication terminal apparatus 10 transmits the instruction "DW" to base station apparatus 20. Base station apparatus 20 receives at time t5 the instruction "UP" transmitted from communication terminal apparatus 10 at time t4, and increases the transmit power to transmit a signal to communication terminal apparatus 10.

[0014] S6 is indicative of SIR of a signal received by communication terminal apparatus 10 at time t6. Since S6 has a value higher than the required SIR value a1, communication terminal apparatus 10 transmits the instruction "DW" to base station apparatus 20. Base station apparatus 20 receives at time t6 the instruction "DW" transmitted from communication terminal apparatus 10 at time t5, and decreases the transmit power to transmit a signal to communication terminal apparatus 10.

[0015] Thus, in communications between communication terminal apparatus 10 and base station apparatus 20, the transmit power control error occurs due to a time delay between the time the transmit power control

information is determined and the time the transmit power is actually updated according to the transmit power control information.

[0016] In the case where the transmit power control information is bit information for instructing an increase or decrease amount in the transmit power, even when the propagation environment of radio signals is static, a time difference arises until the transmit power control information is reflected in a transmit power level, due to a propagation delay of a control signal, and therefore the transmit power level changes more than the increase or decrease amount in the transmit power, and consequently varies widely from the required level. As a result, the SIR value of a signal received by the communication terminal apparatus also varies widely from the required SIR value. In particular, as the delay in the control time is increased, the variation amount in the SIR of the received signal is increased.

[0017] Further, the transmit power control error also occurs under the environment of fading. As a result of the above transmit power variation, the received quality in the communication terminal apparatus deteriorates when the transmit power is controlled below the required power level, while the received qualities in other users may deteriorate when the transmit power is controlled above the required power level.

SUMMARY OF THE INVENTION

[0018] It is an object of the present invention to provide a communication apparatus and transmit power control method that reduce a variation more than an increase or decrease amount in the transmit power in the vicinity of a required level of the transmit power.

[0019] In the closed-loop transmit power control, there occurs a delay between transmitting an instruction on the transmit power control obtained from a received signal quality to a communication partner and reflecting the instruction in the transmit power of the communication partner. Therefore, the present invention achieves the above object by providing a required received quality, which is a criterion for the transmit power control, with a range allowing a transmit power level varying during such a delay to judge, and thereby performing the transmit power control.

[0020] Specifically, the present invention achieves the above object by when the received quality of a received signal is in the range of the required received quality, referring to previous instructions on the transmit power control, instructing the transmit power so that the instructions on the transmit power do not lean in either direction, and thereby reducing the variation in the transmit power due to the propagation delay of a control signal.

[0021] More specifically, the present invention achieves the above object by in the closed-loop transmit power control method, setting a required range from the required level of the received quality, instructing a change opposite to a previously instructed change with respect to the increase or decrease in the transmit power control when the received quality is in the required range, and thereby reducing the variation in the transmit power due to the propagation delay of the control signal.

BRIEF DESCRIPTION OF THE DRAWINGS

[0022] The above and other objects and features of the invention will appear more fully hereinafter from a consideration of the following description taken in connection with the accompanying drawing wherein one example is illustrated by way of example, in which;

FIG.1 is a diagram illustrating a communication aspect of a base station apparatus and a communication terminal apparatus;

FIG.2 is a view showing an example of conventional transmit power control;

FIG.3 is a block diagram illustrating an example of configurations of a communication terminal apparatus and a base station apparatus according to a first embodiment of the present invention;

FIG.4 is a view showing an example of transmit power control between the communication terminal apparatus and base station apparatus according to the first embodiment of the present invention;

FIG.5 is a block diagram illustrating another example of configurations of the communication terminal apparatus and the base station apparatus according to the first embodiment of the present invention;

FIG.6 is a block diagram illustrating another example of configurations of the communication terminal apparatus and the base station apparatus according to the first embodiment of the present invention; and

FIG.7 is a block diagram illustrating an example of configurations of a communication terminal apparatus and a base station apparatus according to a second embodiment of the present invention.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

[0023] Embodiments of the present invention will be described below with reference to accompanying drawings.

(First embodiment)

[0024] The first embodiment of present invention explains a case that in the closed-loop transmit power control, the transmit power control is performed by considering a delay occurring during a period while an instruction on transmit power control obtained from a received signal quality is transmitted to a communication partner and then the instruction is reflected in the transmit power of the communication partner, and providing a required

received quality, which is a criterion for the transmit power control, with a range allowing a transmit power level varying during such a delay to judge.

[0025] FIG.3 is a block diagram illustrating an example of configurations of a communication terminal apparatus and a base station apparatus according to the first embodiment of the present invention.

[0026] Communication terminal apparatus 100 is mainly comprised of antenna 101, receiving section 102, SIR measuring section 103, threshold setting section 104, threshold determining section 105, control bit generating section 106, storage section 107, multiplexing section 108, transmitting section 109 and antenna 110.

[0027] Further, base station apparatus 150 is mainly comprised of antenna 151, receiving section 152, control bit demodulation section 153, transmit power update section 154, transmitting section 155, and antenna 156.

[0028] Generally, a delay occurs during a period while communication terminal apparatus 100 transmits an instruction on the transmit power control to base station apparatus 150 and then base station apparatus 150 transmits a signal with the transmit power with the instruction reflected therein to communication terminal apparatus 100, and therefore the received quality changes by an amount corresponding to the time delay in this loop.

[0029] Hence, the required received quality is provided with a range in advance, and when a received quality is in the range, communication terminal apparatus 100 refers to a previous instruction on the transmit power control, and with respect to an instruction on an increase or decrease in the transmit power, transmits an instruction opposite to the previous instruction to base station apparatus 150.

[0030] Antenna 101 receives a radio signal transmitted from base station apparatus 150 to output to receiving section 102. Receiving section 102 converts the received radio signal into a signal with a baseband frequency to demodulate, and outputs the demodulated signal to SIR measuring section 103.

[0031] SIR measuring section 103 measures SIR (Signal to Interference Ratio) of the demodulated signal to output to threshold determining section 105.

[0032] Threshold setting section 104 calculates an SIR upper limit obtained by adding a predetermined amount to the required SIR value and an SIR lower limit obtained by subtracting a predetermined value from the required SIR value to output to threshold determining section 105.

[0033] When the SIR of the received signal is larger than the SIR upper limit, threshold determining section 105 outputs an instruction for decreasing the transmit power to control bit generating section 106. Further, when the SIR is smaller than the SIR lower limit, threshold determining section 105 outputs an instruction for increasing the transmit power to control bit generating section 106.

[0034] Furthermore, when the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, threshold determining section 105 outputs the determined result to control bit generating section 106.

[0035] According to the determined result in threshold determining section 105, control bit generating section 106 generates a TPC (Transmit Power Control) bit for instructing the transmit power control for a communication partner to output to multiplexing section 108.

[0036] When the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, control bit generating section 106 refers to a last instruction on the transmit power control stored in storage section 107, and outputs an instruction opposite to the last instruction with respect to an increase or decrease in the transmit power to storage section 107 and multiplexing section 108.

[0037] For example, when an instruction for "decreasing the transmit power by 1dB" is last output, control bit generating section 106 next outputs an instruction for "increasing the transmit power by 1dB" to storage section 107 and multiplexing section 108. Meanwhile, when an instruction for "increasing the transmit power by 1dB" is last output, control bit generating section 106 next outputs an instruction for "decreasing the transmit power by 1dB" to storage section 107 and multiplexing section 108.

[0038] Further, when the SIR value is larger than the SIR upper limit, control bit generating section 106 generates a TPC bit indicative of an instruction for decreasing the transmit power to output to storage section 107 and multiplexing section 108. Meanwhile, when the SIR value is smaller than the SIR lower limit, control bit generating section 106 generates a TPC bit indicative of an instruction for increasing the transmit power to output to storage section 107 and multiplexing section 108.

[0039] Storage section 107 stores the instructions on the transmit power control generated in control bit generating section 106.

[0040] Multiplexing section 108 multiplexes transmission data and the TPC bit to output to transmitting section 109. Transmitting section 109 modulates the multiplexed transmission data and TPC bit, converts the resultant signal into a radio signal, and transmits the radio signal through antenna 110.

[0041] Receiving section 152 receives through antenna 151 the radio signal transmitted from communication terminal apparatus 100, and converts the radio signal into a baseband signal to output to control bit demodulation section 153. Control bit demodulation section 153 extracts a transmit power control bit from the received signal to output to transmit power update section 154. Transmit power update section 154 changes the transmit power according to the information contained in the transmit power control bit.

[0042] Thus, communication terminal apparatus 100 measures the SIR of a signal transmitted from base sta-

tion apparatus 150 to judge the transmit power control, and transmits an instruction on the transmit power control to base station apparatus 150. Then, according to the instruction on the transmit power control transmitted from communication terminal apparatus 100, base station apparatus 150 changes the transmit power.

[0043] The operation on the transmit power control will be explained next. FIG.4 is a diagram illustrating an example of the transmit power control between the communication terminal apparatus and base station apparatus according to the first embodiment of the present invention.

[0044] In FIG.4, the abscissa is indicative of time, the ordinate is indicative of SIR, and instructions for communication terminal apparatus 100 to transmit to base station apparatus 150 are indicated at a lower portion in the figure. In FIG.4, the instruction "UP" is an instruction for increasing the transmit power of base station apparatus 150, while the instruction "DW" is another instruction for decreasing the transmit power of base station apparatus 150.

[0045] Communication terminal apparatus 100 sets a required SIR value a1, an SIR lower limit a2, and an SIR upper limit a3.

[0046] Changes in SIR of received signals and instructions on the transmit power control will be explained below with respect to time t1 to t6 in this order.

[0047] S1 is indicative of SIR of a signal received by communication terminal apparatus 100 at time t1. Since S1 has a value lower than the SIR lower limit a2, communication terminal apparatus 100 transmits the instruction "UP" at time t1.

[0048] S2 is indicative of SIR of a signal received by communication terminal apparatus 100 at time t2. Since S2 has a value lower than the SIR lower limit a2, communication terminal apparatus 100 transmits the instruction "UP" at time t2. Base station apparatus 150 receives at time t2 the instruction "UP" transmitted from communication terminal apparatus 100 at time t1, and increases the transmit power to transmit a signal to communication terminal apparatus 100.

[0049] S3 is indicative of SIR of a signal received by communication terminal apparatus 100 at time t3. Since S3 has a value lower than the SIR lower limit a2, communication terminal apparatus 100 transmits the instruction "UP" at time t3. Base station apparatus 150 receives at time t3 the instruction "UP" transmitted from communication terminal apparatus at time t2, and increases the transmit power to transmit a signal to communication terminal apparatus 100.

[0050] S4 is indicative of SIR of a signal received by communication terminal apparatus 100 at time t4. Since S4 has a value equal to or higher than the SIR lower limit a2 and equal to or lower than the SIR upper limit a3, communication terminal apparatus 100 refers to the last transmitted instruction, in other words, the instruction "UP" transmitted at time t3, and transmits the instruction "DW" opposite to the instruction "UP" at time t4.

[0051] Base station apparatus 150 receives at time t4 the instruction "UP" transmitted from communication terminal apparatus at time t3, and increases the transmit power to transmit a signal to communication terminal apparatus 100.

[0052] S5 is indicative of SIR of a signal received by communication terminal apparatus 100 at time t5. Since S5 has a value equal to or higher than the SIR lower limit a2 and equal to or lower than the upper limit a3, communication terminal apparatus 100 refers to the last transmitted instruction, in other words, the instruction "DW" transmitted at time t4, and transmits the instruction "UP" opposite to the instruction "DW" at time t5.

[0053] Base station apparatus 150 receives at time t5 the instruction "DW" transmitted from communication terminal apparatus 100 at time t4, and decreases the transmit power to transmit a signal to communication terminal apparatus 100.

[0054] S6 is indicative of SIR of a signal received by communication terminal apparatus 100 at time t6. Since S6 has a value equal to or higher than the SIR lower limit a2 and equal to or lower than the upper limit a3, communication terminal apparatus 100 refers to the last transmitted instruction, in other words, the instruction "UP" transmitted at time t5, and transmits the instruction "DW" opposite to the instruction "UP" to base station apparatus 150.

[0055] Base station apparatus 150 receives at time t6 the instruction "UP" transmitted from communication terminal apparatus 100 at time t5, and increases the transmit power to transmit a signal to communication terminal apparatus 100.

[0056] Thus, according to the communication terminal apparatus and base station apparatus of the first embodiment of the present invention, in the closed-loop transmit power control, it is possible to reduce the variation more than an increase or decrease amount in the transmit power in the vicinity of the required level of the transmit power, by setting a required range from the required level of the received quality, and when the received quality is in the required range, instructing a change opposite to a previously instructed change with respect to the increase or decrease in the transmit power control.

[0057] In addition, it is preferable that the required range of the received quality is more than or equal to a sum of a transmit power change amount multiplied by a delay time and a transmit power change amount at one time. That is, the required range of the received quality is only required to be more than a amount changing in the transmit power during a delay between transmitting a transmit power control instruction and reflecting the instruction.

[0058] Further, the communication terminal apparatus of this embodiment is capable of changing the threshold into an arbitrary value at an arbitrary timing. FIG.5 is a block diagram illustrating another example of

configurations of the communication terminal apparatus and the base station apparatus in the first embodiment of the present invention. In addition, sections common to those in FIG.3 are assigned the same reference numerals as in FIG.3, and the detailed explanation is omitted.

[0059] Communication terminal apparatus 200 in FIG.5 is different from the communication terminal apparatus in FIG.3 in a point that the apparatus 200 is provided with threshold setting section 201 to change a threshold into an arbitrary value at an arbitrary timing.

[0060] In FIG.5, threshold setting section 201 outputs an instruction for changing a threshold into an arbitrary value at an arbitrary timing to threshold determining section 105. For example, when the signal delay is large and the delay between transmitting a transmit power instruction and reflecting the instruction is increased, threshold setting section 201 outputs an instruction for increasing the SIR upper limit and another instruction for decreasing the SIR lower limit to threshold determining section 105.

[0061] Further, when the signal delay is small and the delay between transmitting a transmit power instruction and reflecting the instruction is decreased, threshold setting section 201 outputs an instruction for decreasing the SIR upper limit and another instruction for increasing the SIR lower limit to threshold determining section 105.

[0062] According to the instructions output from threshold setting section 201, threshold determining section 105 changes the SIR upper limit and SIR lower limit to compare with the SIR of a received signal.

[0063] Thus, communication terminal apparatus 200 of this embodiment changes the SIR upper limit and SIR lower limit corresponding to a delay in the propagation path, and thereby makes a range from the SIR upper limit to the SIR lower limit more than a change amount in the transmit power control due to the propagation delay. As a result, when the SIR value is in the vicinity of the required SIR value, it is possible to prevent the transmit power from being changed more than a change in the transmit power during a period from the time the received quality is compared to the time the instruction on the transmit power is reflected.

[0064] Further, the communication terminal apparatus of this embodiment is capable of changing the SIR upper limit and SIR lower limit according to a change of the required SIR value. FIG.6 is a block diagram illustrating another example of configurations of the communication terminal apparatus and the base station apparatus in the first embodiment of the present invention. In addition, sections common to those in FIG.3 are assigned the same reference numerals as in FIG.3, and the detailed explanation is omitted.

[0065] Communication terminal apparatus 300 in FIG.6 is different from the communication terminal apparatus in FIG.3 in a point that the apparatus 300 is provided with threshold setting section 301 to change the SIR upper limit and SIR lower limit according to a change of the required SIR value.

[0066] In FIG.6, when the required SIR value is changed, threshold setting section 301 outputs an instruction for changing the SIR upper limit and the SIR lower limit by the same increase or decrease amount as in changing the required SIR value. According to the instruction output from threshold setting section 301, threshold determining section 105 changes the SIR upper limit and the SIR lower limit to compare with SIR of a received signal.

[0067] Further, during a period or a plurality of periods of measuring SIR immediately after changing the required SIR value, the threshold determining section 301 instructs to decrease the transmit power when the measured SIR value is more than or equal to the required SIR value, while instructing to increase the transmit power when the measured SIR value is less than the required SIR value.

[0068] Thus, communication terminal apparatus 300 of this embodiment changes the SIR upper limit and the SIR lower limit by the same increase or decrease amount as in changing the required SIR value, whereby the apparatus 300 is capable of reducing a variation more than an increase or decrease amount in the transmit power in the vicinity of the required level of the transmit power even when the required SIR value is changed.

(Second embodiment)

[0069] FIG.7 is a block diagram illustrating an example of configurations of a communication terminal apparatus and a base station apparatus according to the second embodiment of the present invention.

[0070] In FIG.7, communication terminal apparatus 400 is mainly comprised of antenna 401, receiving section 402, SIR measuring section 403, threshold setting section 404, threshold determining section 405, control bit generating section 406, storage section 407, multiplexing section 408, control bit demodulation section 409, transmit power update section 410, transmitting section 411 and antenna 412.

[0071] Further, base station apparatus 450 is mainly comprised of antenna 451, receiving section 452, SIR measuring section 453, threshold setting section 454, threshold determining section 455, control bit generating section 456, storage section 457, multiplexing section 458, control bit demodulation section 459, transmit power update section 460, transmitting section 461, and antenna 462.

[0072] Antenna 401 receives a radio signal transmitted from base station apparatus 450 to output to receiving section 402. Receiving section 402 converts the received radio signal into a signal with a baseband frequency to demodulate, and outputs the demodulated signal to SIR measuring section 403 and control bit demodulation section 409.

[0073] SIR measuring section 403 measures SIR of the demodulated signal to output to threshold determin-

ing section 405. Threshold setting section 404 calculates an SIR upper limit obtained by adding a predetermined amount to the required SIR value and an SIR lower limit obtained by subtracting a predetermined value from the required SIR value to output to threshold determining section 405.

[0074] When the SIR of the received signal is larger than the SIR upper limit, threshold determining section 405 outputs an instruction for decreasing the transmit power to control bit generating section 406. Further, when the SIR is smaller than the SIR lower limit, threshold determining section 405 outputs an instruction for increasing the transmit power to control bit generating section 406.

[0075] Furthermore, when the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, threshold determining section 405 outputs the determined result to control bit generating section 406. According to the determined result in threshold determining section 405, control bit generating section 406 generates a TPC bit for instructing the transmit power control for a communication partner to output to multiplexing section 408.

[0076] When the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, control bit generating section 406 refers to a last instruction on the transmit power control stored in storage section 407, and outputs an instruction opposite to the last instruction with respect to an increase or decrease in the transmit power to storage section 407 and multiplexing section 408. The specific operation of control bit generating section 406 is the same as in control bit generating section 106 of the first embodiment.

[0077] Further, when the SIR of the received signal is larger than the SIR upper limit, control bit generating section 406 generates a TPC bit indicative of an instruction for decreasing the transmit power to output to storage section 407 and multiplexing section 408. Meanwhile, when the SIR of the received signal is smaller than the SIR lower limit, control bit generating section 406 generates a TPC bit indicative of an instruction for increasing the transmit power to output to storage section 407 and multiplexing section 408.

[0078] Storage section 407 stores the instructions on the transmit power control generated in control bit generating section 406. Multiplexing section 408 multiplexes transmission data and the TPC bit to output to transmitting section 411.

[0079] Control bit demodulation section 409 extracts a transmit power control bit from the received signal to output to transmit power update section 410. Transmit power update section 410 changes the transmit power according to the information contained in the transmit power control bit. Transmitting section 411 modulates the multiplexed transmission data and TPC bit, converts the resultant signal to a radio signal, and transmits the radio signal through antenna 412.

[0080] Antenna 451 receives the radio signal transmitted from communication terminal apparatus 400 to output to receiving section 452. Receiving section 452 converts the received radio signal into a signal with a baseband frequency to demodulate, and outputs the demodulated signal to SIR measuring section 453 and control bit demodulation section 459.

[0081] SIR measuring section 453 measures SIR of the demodulated signal to output to threshold determining section 455. Threshold setting section 454 calculates an SIR upper limit obtained by adding a predetermined amount to the required SIR value and an SIR lower limit obtained by subtracting a predetermined value from the required SIR value to output to threshold determining section 455.

[0082] When the SIR of the received signal is larger than the SIR lower limit, threshold determining section 455 outputs an instruction for decreasing the transmit power to control bit generating section 456. Further, when the SIR is smaller than the SIR upper limit, threshold determining section 455 outputs an instruction for increasing the transmit power to control bit generating section 456.

[0083] Furthermore, when the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, threshold determining section 455 outputs the determined result to control bit generating section 456. According to the determined result in threshold determining section 455, control bit generating section 456 generates a TPC bit for instructing the transmit power control for a communication partner to output to multiplexing section 458.

[0084] When the SIR of the received signal is equal to or smaller than the SIR upper limit and equal to or larger than the SIR lower limit, control bit generating section 456 refers to a last instruction on the transmit power control stored in storage section 457, and outputs an instruction opposite to the last instruction with respect to an increase or decrease in the transmit power to storage section 457 and multiplexing section 458. The specific operation of control bit generating section 456 is the same as in control bit generating section 106 of the first embodiment.

[0085] Further, when the SIR is larger than the SIR upper limit, control bit generating section 456 generates a TPC bit indicative of an instruction for decreasing the transmit power to output to storage section 457 and multiplexing section 458. Meanwhile, when the SIR is smaller than the SIR lower limit, control bit generating section 456 generates a TPC bit indicative of an instruction for increasing the transmit power to output to storage section 457 and multiplexing section 458.

[0086] Storage section 457 stores the instructions on the transmit power control generated in control bit generating section 456. Multiplexing section 458 multiplexes transmission data and the TPC bit to output to transmitting section 461.

[0087] Control bit demodulation section 459 extracts

a transmit power control bit from the received signal to output to transmit power update section 460. Transmit power update section 460 changes the transmit power according to the information contained in the transmit power control bit. Transmitting section 461 modulates the multiplexed transmission data and TPC bit, converts the resultant signal to a radio signal, and transmits the radio signal through antenna 462.

[0088] Thus, communication terminal apparatus 400 measures the SIR of a signal transmitted from base station apparatus 450 to judge the transmit power control, and transmits an instruction on the transmit power control to base station apparatus 450. Then, according to the instruction on the transmit power control transmitted from communication terminal apparatus 400, base station apparatus 450 changes the transmit power.

[0089] Further, base station apparatus 450 measures the SIR of a signal transmitted from communication terminal apparatus 400 to judge the transmit power control, and transmits an instruction on the transmit power control to communication terminal apparatus 400. Then, according to the instruction on the transmit power control transmitted from base station apparatus 450, communication terminal apparatus 400 changes the transmit power.

[0090] Thus, according to communication terminal apparatus 400 and base station apparatus 450 of the second embodiment of the present invention, in the closed-loop transmit power control, it is possible to reduce the variation more than an increase or decrease amount in the transmit power in the vicinity of the required level of the transmit power, by setting a required range from the required level of the received quality, and when the received quality is in the required range, instructing a change opposite to a previously instructed change with respect to the increase or decrease in the transmit power control.

[0091] In an apparatus of the communication partner, the above processing is performed in the same way, whereby the transmit power control information to be transmitted and a transmit power level of a signal to be transmitted are determined.

[0092] In addition, while this embodiment describes about a communication terminal apparatus and/or base station apparatus, the present invention is not limited to those, and any communication apparatuses are applicable that control transmit power of a communication partner from a received quality of a signal transmitted from the communication partner.

[0093] Further, while the embodiments of the present invention explain a case that the SIR of a received signal is compared with the required SIR, the present invention is not limited to the above case, and any values are applicable that are indicative of a received quality.

[0094] The transmit power control apparatus in the CDMA radio communication as described above is capable of reducing a variation in the measured SIR value in the vicinity of the required SIR value due to a delay in control time. As a result, the apparatus is capable of properly controlling the transmit power level to an extent enabling a predetermined received quality to be maintained and further of decreasing the interference in other users.

[0095] As this invention may be embodied in several forms without departing from the spirit of essential characteristics thereof, the present embodiment is therefore illustrative and not restrictive, since the scope of the invention is defined by the appended claims rather than by the description preceding them, and all changes that fall within meets and bounds of the claims, or equivalence of such meets and bounds are therefore intended to embraced by the claims.

[0096] The present invention is not limited to the above described embodiments, and various variations and modifications may be possible without departing from the scope of the present invention.

[0097] This application is based on the Japanese Patent Application No.2000-057195 filed on March 2, 2000, entire content of which is expressly incorporated by reference herein.

Claims

1. A radio communication apparatus comprising:

> measuring means (103) for measuring a received quality of a signal transmitted from a communication partner;
> instructing means (105,106) for performing an instruction for changing transmit power to said communication partner from a result obtained by comparing the received quality of the signal with a required received quality;
> storage means (107) for storing the instruction for changing transmit power; and
> transmitting means (109) for transmitting the instruction for changing transmit power to said communication partner,

wherein said instructing means provides a value of the required received quality with a range allowing a change in the transmit power caused by a propagation delay of a transmit power control signal to perform the instruction for changing the transmit power.

2. The radio communication apparatus according to claim 1, wherein said instructing means (105,106) performs the instruction for decreasing the transmit power to said communication partner when the received quality of the signal is higher than an upper limit of a predetermined range, further performs the instruction for increasing the transmit power to said communication partner when the received quality of the signal is lower than a lower limit of the prede-

termined range, and refers to a previous instruction for changing the transmit power stored in said storage means (107) to perform the instruction for changing the transmit power so as to maintain the received quality of the signal in the predetermined range when the received quality of the signal is in the predetermined range.

3. The radio communication apparatus according to claim 2, wherein said instructing means (105,106) refers to the previous instruction for changing the transmit power stored in said storage means (107) when the received quality of the signal is in the predetermined range, and when a last instruction for changing transmit power is indicative of increasing the transmit power, generates the instruction for decreasing the transmit power, while when the last instruction for changing transmit power is indicative of decreasing the transmit power, generating the instruction for increasing the transmit power.

4. The radio communication apparatus according to claim 3, further comprising:

setting means (104) for setting the predetermined range with a predetermined variation range from the required received quality.

5. The radio communication apparatus according to claim 1, wherein said instructing means (105,106) changes the range into an arbitrary range at an arbitrary timing.

6. The radio communication apparatus according to claim 1, wherein when the required received quality is changed, said instructing means (105,106) performs the instruction for decreasing the transmit power when the received quality of the signal is higher than the required received quality, while performing the instruction for increasing the transmit power when the received quality of the signal is lower than the required received quality, not depending on whether the received quality of the signal is in the range during one or a plurality of measuring periods immediately after changing the received quality.

7. A mobile station apparatus having a radio communication apparatus, said radio communication apparatus comprising:

measuring means (103) for measuring a received quality of a signal transmitted from a communication partner;
instructing means (105,106) for generating an instruction for changing transmit power to said communication partner from a result obtained by comparing the received quality of the signal with a required received quality;
storage means (107) for storing the instruction for changing transmit power; and
transmitting means (109) for transmitting the instruction for changing transmit power to said communication partner,

wherein said instructing means performs the instruction for decreasing the transmit power to said communication partner when the received quality of the signal is higher than an upper limit of a predetermined range, further performs the instruction for increasing the transmit power to said communication partner when the received quality of the signal is lower than a lower limit of the predetermined range, and refers to a previous instruction for changing the transmit power stored in said storage means to perform the instruction for changing the transmit power so as to maintain the received quality of the signal in the predetermined range when the received quality of the signal is in the predetermined range.

8. A base station apparatus having a radio communication apparatus, said radio communication apparatus comprising:

measuring means (103) for measuring a received quality of a signal transmitted from a communication partner;
instructing means (105,106) for generating an instruction for changing transmit power to said communication partner from a result obtained by comparing the received quality of the signal with a required received quality;
storage means (107) for storing the instruction for changing transmit power; and
transmitting means (109) for transmitting the instruction for changing transmit power to said communication partner,

wherein said instructing means performs the instruction for decreasing the transmit power to said communication partner when the received quality of the signal is higher than an upper limit of a predetermined range, further performs the instruction for increasing the transmit power to said communication partner when the received quality of the signal is lower than a lower limit of the predetermined range, and refers to a previous instruction for changing the transmit power stored in said storage means to perform the instruction for changing the transmit power so as to maintain the received quality of the signal in the predetermined range when the received quality of the signal is in the predetermined range.

9. A transmit power control method, comprising:

measuring a received quality of a signal transmitted from a communication partner;
comparing the received quality of the signal with a required received quality, and based on the compared result, performing an instruction for changing transmit power to said communication partner
storing the instruction for changing transmit power; and
transmitting the instruction for decreasing the transmit power to said communication partner when the received quality of the signal is higher than an upper limit of a predetermined range, further transmitting the instruction for increasing the transmit power to said communication partner when the received quality of the signal is lower than a lower limit of the predetermined range, and referring to a stored previous instruction for changing the transmit power to transmit the instruction for changing the transmit power such that the received quality of the signal is maintained in the predetermined range to said communication partner when the received quality of the signal is in the predetermined range.

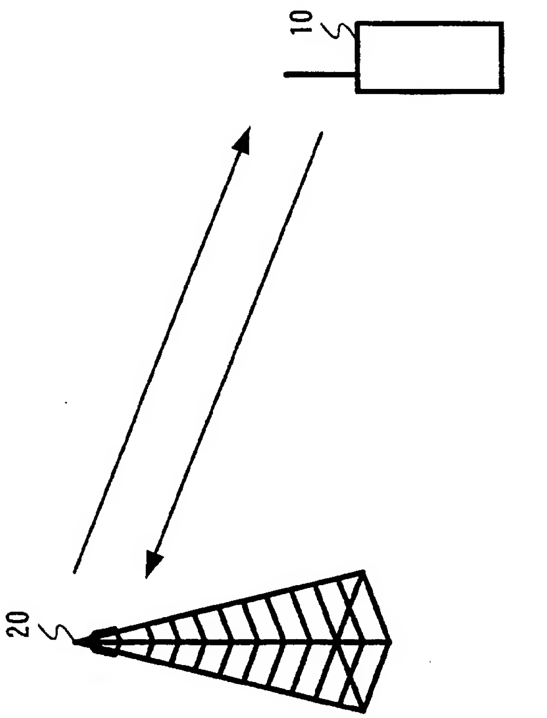
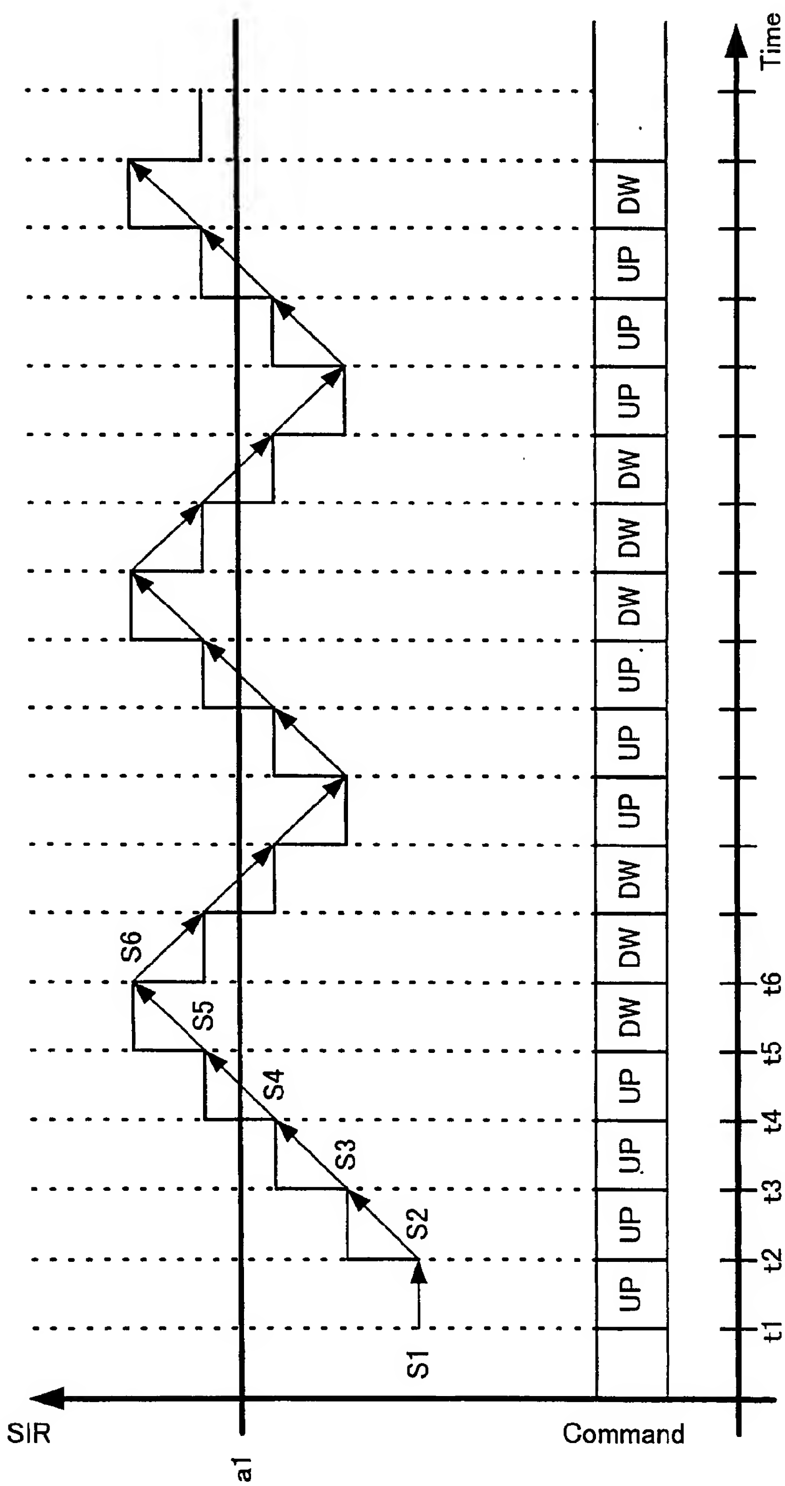


FIG. 1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a1
Command
Time
S1
S2
S3
S4
S5
S6
t1
t2
t3
t4
t5
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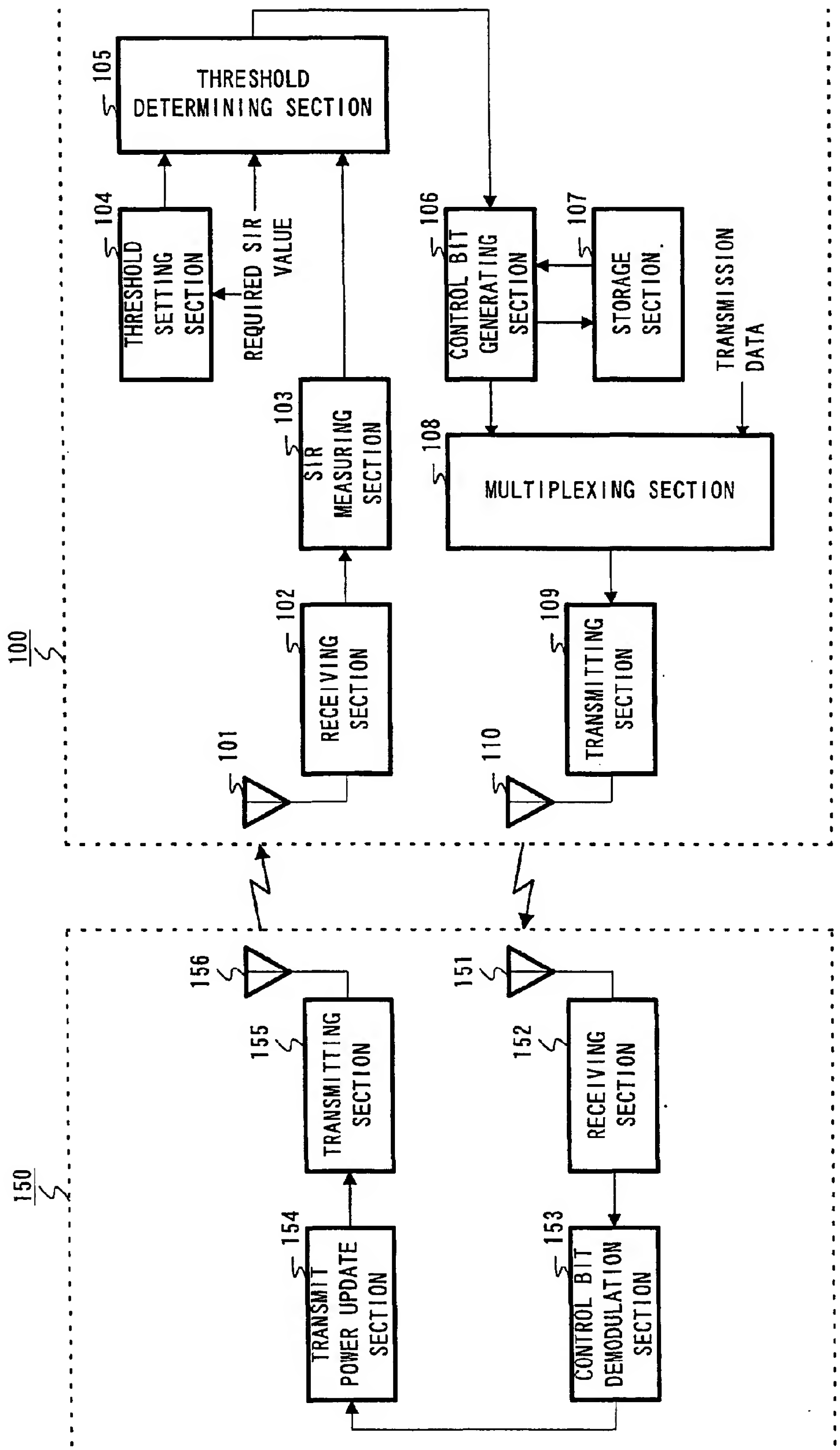
UP
UP
UP
DW

FIG. 2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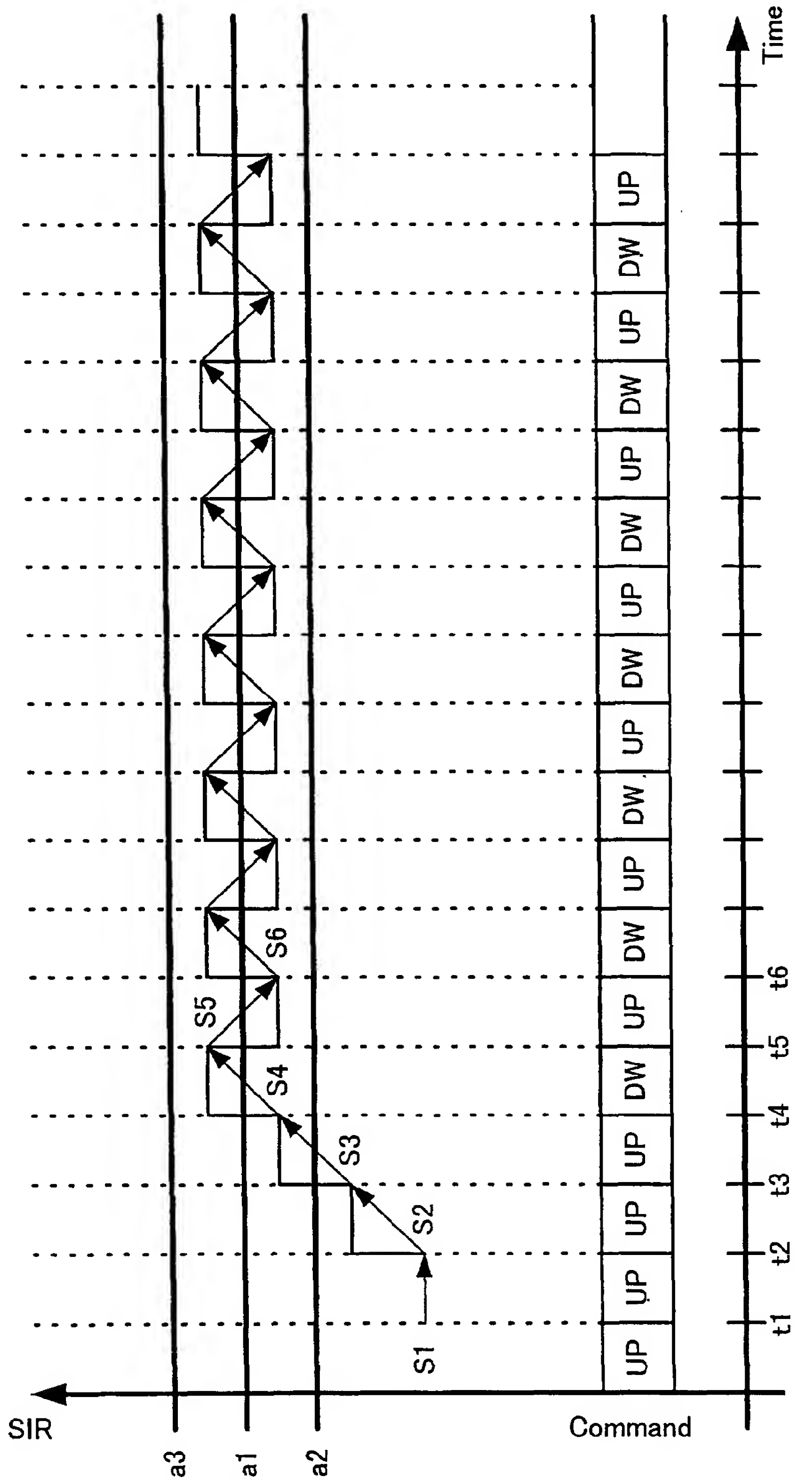
RECEIVING SECTION
102
SIR MEASURING SECTION
103
THRESHOLD SETTING SECTION
104
REQUIRED SIR VALUE
THRESHOLD DETERMINING SECTION
105
CONTROL BIT GENERATING SECTION
106
STORAGE SECTION
107
TRANSMISSION DATA
MULTIPLEXING SECTION
108
TRANSMITTING SECTION
109
110
150
151
RECEIVING SECTION
152
CONTROL BIT DEMODULATION SECTION
153
TRANSMIT POWER UPDATE SECTION
154
TRANSMITTING SECTION
155
156

FIG. 3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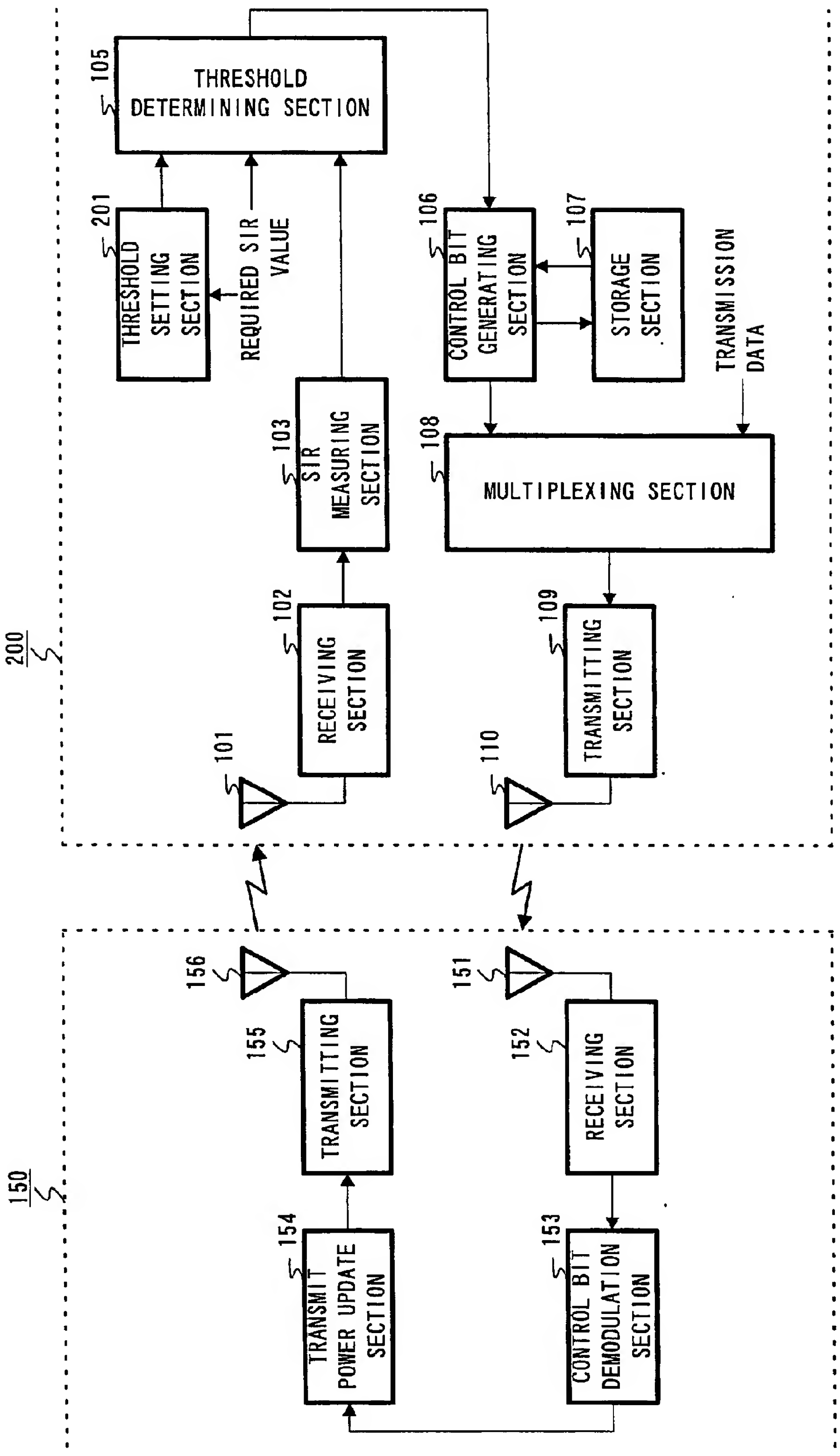
a3
a1
a2
S1
S2
S3
S4
S5
S6
Command
UP
UP
UP
UP
DW
UP
DW
UP
DW
UP
DW
UP
DW
UP
DW
UP
Time
t1
t2
t3
t4
t5
t6

FIG. 4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THRESHOLD SETTING SECTION
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CONTROL BIT GENERATING SECTION
106
STORAGE SECTION
107
TRANSMISSION DATA
MULTIPLEXING SECTION
108
TRANSMITTING SECTION
109
110
150
156
TRANSMITTING SECTION
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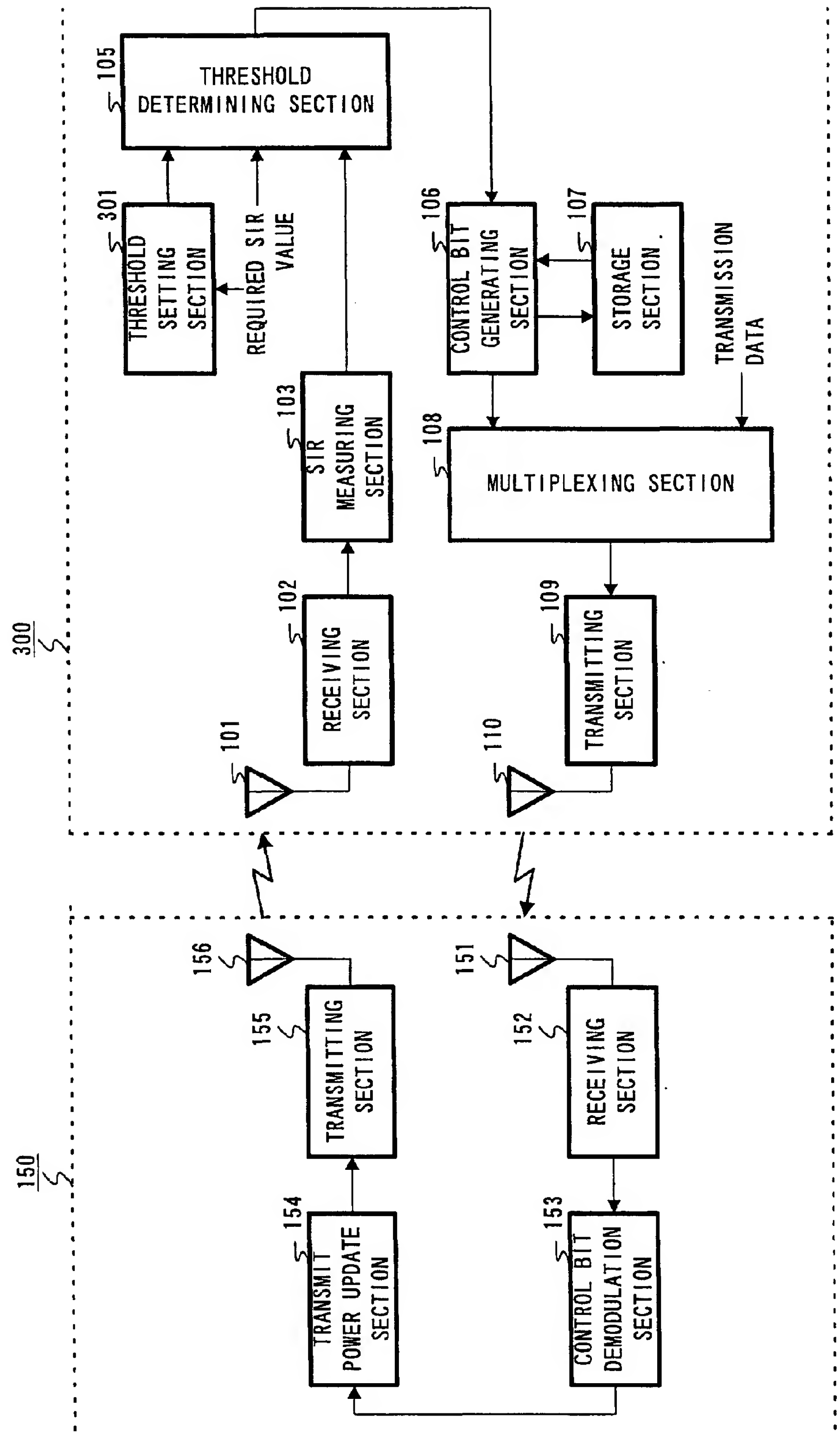
TRANSMIT POWER UPDATE SECTION
154
151
RECEIVING SECTION
152
CONTROL BIT DEMODULATION SECTION
153

FIG. 5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105
THRESHOLD DETERMINING SECTION
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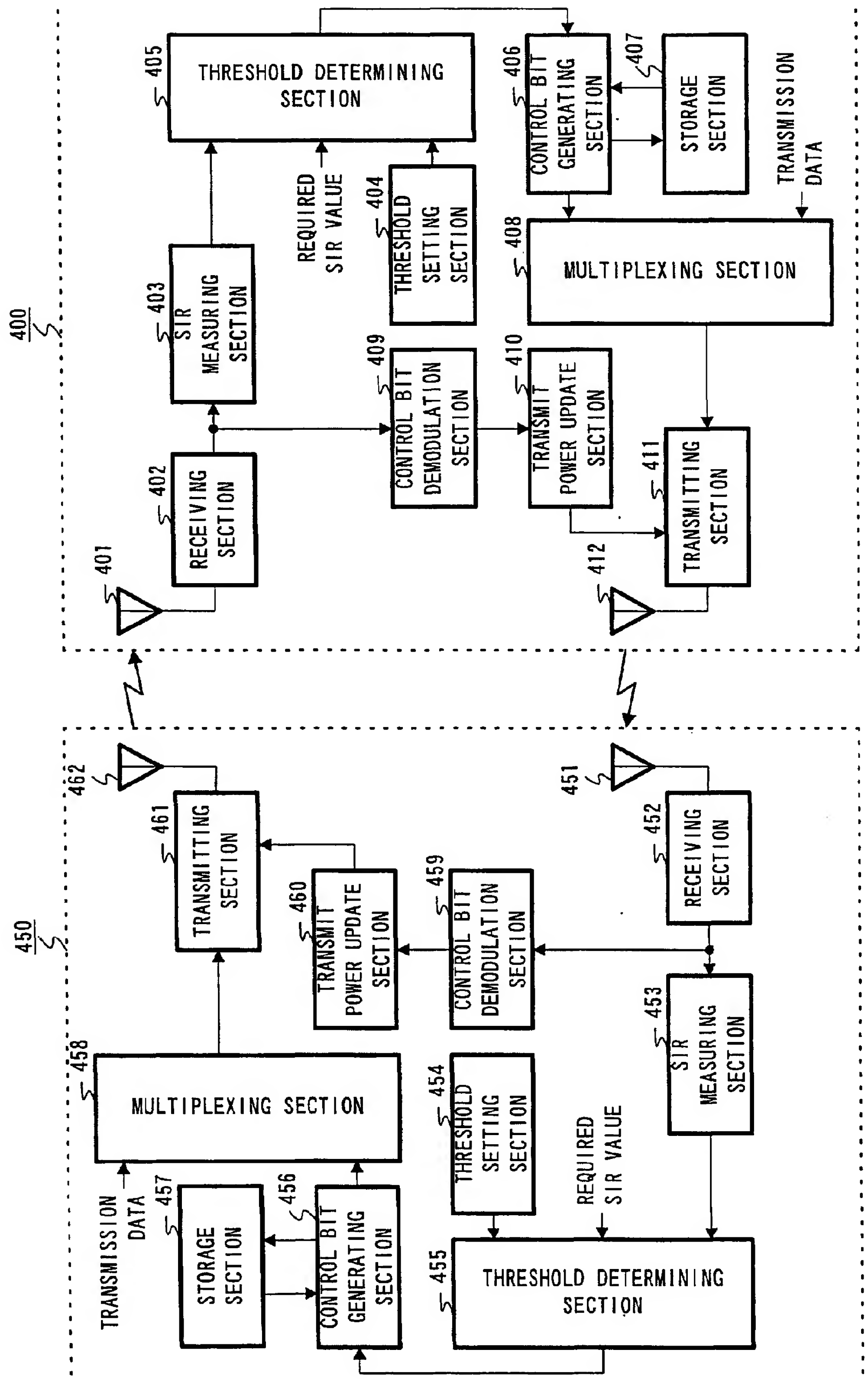
THRESHOLD SETTING SECTION
REQUIRED SIR VALUE
106
CONTROL BIT GENERATING SECTION
107
STORAGE SECTION
TRANSMISSION DATA
103
SIR MEASURING SECTION
108
MULTIPLEXING SECTION
102
RECEIVING SECTION
109
TRANSMITTING SECTION
101
110
150
156
155
TRANSMITTING SECTION
151
152
RECEIVING SECTION
154
TRANSMIT POWER UPDATE SECTION
153
CONTROL BIT DEMODULATION SECTION

FIG. 6


400
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SIR MEASURING SECTION
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THRESHOLD SETTING SECTION
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THRESHOLD DETERMINING SECTION
405
REQUIRED SIR VALUE
CONTROL BIT GENERATING SECTION
406
STORAGE SECTION
407
MULTIPLEXING SECTION
408
TRANSMISSION DATA
CONTROL BIT DEMODULATION SECTION
409
TRANSMIT POWER UPDATE SECTION
410
TRANSMITTING SECTION
411
412
450
451
RECEIVING SECTION
452
SIR MEASURING SECTION
453
THRESHOLD SETTING SECTION
454
THRESHOLD DETERMINING SECTION
455
REQUIRED SIR VALUE
CONTROL BIT GENERATING SECTION
456
STORAGE SECTION
457
MULTIPLEXING SECTION
458
TRANSMISSION DATA
CONTROL BIT DEMODULATION SECTION
459
TRANSMIT POWER UPDATE SECTION
460
TRANSMITTING SECTION
461
462

FIG. 7

(19) RU (11) 2 157 592 (13) C2

(51) МПК⁷ H 04 B 7/26

РОССИЙСКОЕ АГЕНТСТВО
ПО ПАТЕНТАМ И ТОВАРНЫМ ЗНАКАМ

(12) ОПИСАНИЕ ИЗОБРЕТЕНИЯ К ПАТЕНТУ РОССИЙСКОЙ ФЕДЕРАЦИИ

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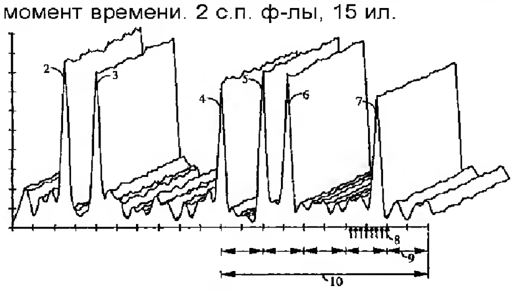
RU 2157592 C2

(54) СПОСОБ ПРИЕМА И ПОИСКА СИГНАЛА, ПЕРЕДАВАЕМОГО ПАКЕТАМИ

(57)
Интегральный поисковый процессор, используемый в модеме для системы связи с расширенным спектром, буферизирует в буфере выборки принимаемых сигналов и использует процессор преобразования с квантованием времени, работающий с последовательными сдвигами по отношению к буферу. Поисковый процессор осуществляет пошаговый автономный поиск, конфигурируемый микропроцессором, в котором определен набор параметров поиска, который может включать группу антенн для поиска, начальный сдвиг и ширину поискового окна для проведения поиска и количество символов Уолша для накопления результатов на каждом сдвиге. Поисковый процессор вычисляет энергию корреляции на каждом сдвиге и представляет итоговый список наилучших трасс распространения сигналов, обнаруженных в процессе поиска, для использования их при расширении элементов демодуляции, что является техническим результатом. Поиск выполняется линейным образом независимо от вероятности того, что искомый сигнал передавался в любой данный момент времени. 2 с.п. ф-лы, 15 ил.



Фиг.1

RU 2157592 C2

(19) RU (11) 2 157 592 (13) C2

(51) Int. Cl.7 **H 04 B 7/26**

RUSSIAN AGENCY
FOR PATENTS AND TRADEMARKS

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(73) Proprietor:
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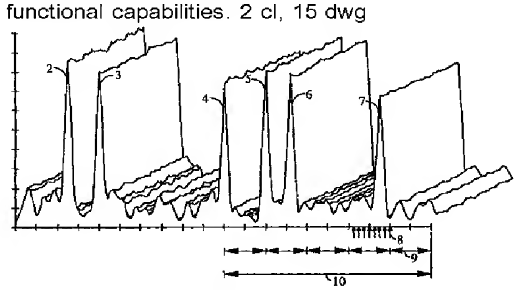
(54) METHOD FOR RECEPTION AND SEARCH OF PACKET- SWITCHED SIGNAL

(57) Abstract:
FIELD: communication equipment. SUBSTANCE: integral search processor, which is used in modem of extended spectrum communication system, provides buffering of samples of received signals in buffer and uses conversion processor, which uses time sampling and operates in sequential steps with respect to buffer. Search processor achieves step-by-step autonomous search, which is configured by microprocessor, which defines set of search parameters, including group of antennas for search, initial shift and search window width, as well as number of Walsh characters for accumulation of results for each shift. Search processor calculates correlation power for each shift and produces final list of best possible signal spreading tracks, which are detected upon search, in order to use them for extension of demodulation elements. Search is linear and independent from probability of the fact that target signal was sent in any arbitrary time moment. EFFECT: increased functional capabilities. 2 cl, 15 dwg



Фиг.1

RU 2157592 C2

Область техники

Настоящее изобретение относится к системам связи с расширенным спектром, более конкретно к обработке сигнала в сотовой телефонной системе связи.

Описание известного уровня техники

В радиотелефонных системах связи, таких как сотовые телефонные системы, системы персональной связи, и локальные замкнутые системы радиосвязи, многие пользователи осуществляют связь по радиоканалу для подсоединения к проводным телефонным системам. Связь по радиоканалу может быть одним из множества методов множественного доступа, обеспечивающих возможность большому числу абонентов использовать ограниченный частотный спектр. Эти методы множественного доступа включают множественный доступ с временным разделением /МДВР/ каналов, множественный доступ с частотным разделением /МДЧР/ и множественный доступ с кодовым разделением /МДКР/. Метод МДКР имеет множество преимуществ, причем типовая система с МДКР раскрыта в Патенте США N 4901307, от 13 февраля 1990 на "Систему связи множественного доступа с расширенным спектром, использующую спутниковые или наземные ретрансляторы", переуступленном правопреемнику настоящего изобретения.

В упомянутом патенте раскрыт способ множественного доступа, в котором большое количество мобильных пользователей телефонной системы, каждый из которых имеет приемопередатчик, осуществляют связь через спутниковые ретрансляторы или наземные базовые станции с использованием сигналов связи в режиме МДКР с расширенным спектром. При осуществлении связи в режиме МДКР частотный спектр может использоваться многократно, что позволяет увеличить пропускную способность системы для пользователей.

Способы модуляции МДКР, раскрытые в Патенте США N 4901307, имеют множество преимуществ по сравнению со способами узкополосной модуляции, применяемыми в системах связи, использующих спутниковые и наземные каналы. В наземных каналах возникают специфические требования к системе связи, в частности, по отношению к многолучевым сигналам. Применение способов МДКР позволяет решить специфические проблемы, связанные с использованием наземного канала, путем преодоления отрицательного влияния многолучевого распространения, например, замирания, не теряя в то же время их преимуществ.

Способ МДКР, раскрытый в Патенте США N 4901307, предусматривает использование когерентной модуляции и демодуляции для обоих направлений линии связи при осуществлении связи между удаленным устройством радиосвязи и спутником. В нем соответственно раскрывается использование несущего пилот-сигнала в качестве когерентного фазового опорного сигнала для линии связи "спутник-удаленное устройство радиосвязи" и линии связи "базовая станция-удаленное устройство радиосвязи". Однако в наземном оборудовании сотовой связи наличие весьма сильного многолучевого затухания, приводящего к искажению фазы каналом, а также уровень мощности, необходимый для передачи несущего пилот-сигнала от удаленного устройства радиосвязи, затрудняет использование способа когерентной модуляции для линии связи "удаленное устройство радиосвязи-базовая станция". В Патенте США N 5103549 на "Систему и способ формирования сигналов в сотовой телефонной системе МДКР" от 25 июня 1990 года, переуступленном правопреемнику настоящего изобретения, предлагается средство для преодоления отрицательного влияния многолучевого распространения в канале связи "удаленное устройство радиосвязи-базовая станция" с использованием способов некогерентной модуляции и демодуляции.

В сотовой телефонной системе с МДКР одна и та же полоса частот может быть использована для связи со всеми базовыми станциями. В приемнике базовой станции разделяемые сигналы многолучевого распространения, например, проступающие по трассе от абонентского пункта и по другой трассе после отражения от здания, могут быть объединены при обработке с разнесением для улучшения характеристик модема. Свойства сигналов МДКР, обеспечивающие выигрыш при обработке, также используются для различения сигналов, занимающих один и тот же частотный диапазон. Кроме того высокоскоростная псевдошумовая модуляция всегда обеспечивает возможность разделения различных трасс распространения одного и того же сигнала, при условии, если разница в задержках по трассам распространения превышает длительность элемента псевдошумового кода. Если в системе с МДКР используется частота следования элементов псевдошумового кода порядка 1 МГц, то выигрыш при обработке сигнала расширенного спектра, равный отношению расширенной полосы к частоте данных в системе, может быть получен для всех трасс распространения, имеющих задержки, отличающиеся более чем на одну микросекунду. Разница в задержках на трассе распространения в одну микросекунду соответствует разнице в расстояниях примерно 300 метров. Обычно городская среда обеспечивает разность задержек по трассам распространения сигналов, превышающую одну микросекунду.

Свойства наземного канала, связанные с многолучевым распространением, приводят к тому, что в приемник сигналы приходят по нескольким различным трассам распространения. Одной из характеристик канала с многолучевым распространением является разброс по времени, возникающий в сигнале, передаваемом через такой канал. Например, если по каналу с многолучевым распространением передается идеальный импульс, то принимаемый сигнал появляется в виде последовательности импульсов. Другой характерной особенностью канала с многолучевым распространением является то, что каждая трасса распространения сигнала может давать разный коэффициент затухания. Например, если по каналу с многолучевым распространением передается идеальный импульс, то каждый импульс принимаемой импульсной

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последовательности обычно имеет уровень, отличный от уровня других принимаемых импульсов. Еще одной характерной особенностью канала с многолучевым распространением является то, что каждая трасса распространения дает разную фазу сигнала. Например, если по каналу с многолучевым распространением передается идеальный импульс, то каждый импульс принимаемой последовательности обычно имеет фазу, отличающуюся от фазы других принимаемых импульсов.

При передаче по радиоканалу многолучевое распространение возникает благодаря отражению сигнала от препятствий, образуемых зданиями, деревьями, автомобилями и людьми. В общем случае радиоканал представляет собой нестационарный канал с многолучевым распространением из-за относительного перемещения объектов, создающих многолучевое распространение. Например, если по нестационарному каналу с многолучевым распространением передается идеальный импульс, то в принимаемой последовательности импульсов будет изменяться его положение во времени, затухание и фаза в функции времени передачи идеального импульса.

Многолучевое распространение может вызвать замирание сигнала в канале. Замирание является результатом характеристик фазирования канала с многолучевым распространением. Замирание появляется, когда векторы множества лучей суммируются неблагоприятным образом, образуя в результате принимаемый сигнал, меньший, чем любой отдельно взятый вектор. Например, если по каналу с многолучевым распространением передается гармонический сигнал, характеризуемый двумя трассами распространения, причем первая трасса имеет коэффициент ослабления X дБ, время задержки σ с фазовым сдвигом θ радиан, а вторая трасса имеет коэффициент ослабления X дБ, время задержки $\theta+\pi$ с фазовым сдвигом σ радиан, то на выходе канала принятый сигнал будет отсутствовать.

В системах узкополосной модуляции, таких как аналоговая частотная модуляция, используемая в известных радиотелефонных системах, наличие многолучевого распространения в радиоканале приводит к существенному многолучевому замиранию. Однако, как было отмечено выше в случае широкополосной системы МДКР, в процессе демодуляции могут быть выделены разные трассы распространения. Такое разделение не только значительно уменьшает отрицательное воздействие замирания, но и дает преимущества, связанные с использованием системы МДКР.

Разнесение - один из способов уменьшения отрицательных влияний замирания. Следовательно, желательно обеспечить некоторую форму разнесения, что позволит уменьшить замирание в системе. Существуют три основных вида разнесения: временное разнесение, частотное разнесение и пространственное разнесение или разнесение по трассе распространения.

Временное разнесение обеспечивается с использованием повторения временного перемежения и кодирования с обнаружением и исправлением ошибок, которое вводит избыточность. В системе, использующей настоящее изобретение, можно применять любой из этих способов в качестве формы временного разнесения.

МДКР с присущей данному методу широкополосностью обеспечивает некоторую форму частотного разнесения путем распределения энергии сигнала в широкой полосе рабочих частот. Поэтому частотно-избирательное замирание проявляется только на небольшой части полосы рабочих частот сигнала МДКР.

Пространственное разнесение и разнесение по трассе распространения обеспечивается посредством подачи сигнала с многолучевым распространением одновременно по нескольким линиям связи от удаленного устройства радиосвязи через две или более базовые станции и путем использования двух или более пространственно разнесенных антенных элементов на одной базовой станции. Кроме того разнесение по трассам распространения может быть получено путем использования среды многолучевого распространения посредством обработки расширенного спектра, что позволяет осуществлять прием и раздельную обработку сигнала, поступающего с различными задержками распространения, как было описано выше. Примеры разнесения по трассам распространения приведены в Патенте США N 5101501 на "Программируемое переключение связи в сотовой телефонной системе МДКР" от 21 марта 1992 года и Патенте США N 5109390 на "Приемник разнесенного приема в сотовой телефонной системе МДКР" от 8 октября 1991, переуступленных правопреемнику настоящего изобретения.

Отрицательное влияние замирания может быть до некоторой степени скомпенсировано в системе с МДКР посредством регулирования мощности передатчика. Система для управления мощностью базовой станции и удаленного модуля раскрыта в Патенте США N 5056109 на "Способ и устройство для управления передаваемой мощностью в сотовой мобильной телефонной системе МДКР" от 8 октября 1991 года, переуступленном правопреемнику настоящего изобретения.

Способ МДКР, раскрытый в Патенте США N 4901307, предусматривает использование относительно длинных псевдоожиженных последовательностей /ПШП/, причем пользователю каждого удаленного устройства радиосвязи выделяется своя, отличная от других ПШП. Взаимная корреляция между различными ПШП и автокорреляция ПШП для всех временных сдвигов, отличных от нуля, имеют среднее значение, близкое к нулю, что позволяет различать сигналы различных пользователей при приеме. (Для получения нулевого среднего значения для автокорреляции и взаимной корреляции, требуется, чтобы логический "0" принял значение "1", а логическая "1" приняла значение "-1" или на исходное отображение логических уровней).

Однако такие псевдошумовые сигналы не ортогональны. Хотя взаимная корреляция по существу имеет нулевое среднее значение на всей длине последовательности, однако в течение короткого временного интервала, такого как время информационного бита,

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взаимная корреляция является случайной переменной с биномиальным распределением. При этом сигналы взаимодействуют друг с другом в основном таким же образом, как если бы они представляли собой широкополосный шум с гауссовым распределением с той же самой спектральной плотностью мощности. Таким образом, сигналы других пользователей или шум от взаимных помех в конечном счете ограничивают достигаемую пропускную способность.

Специалистам хорошо известно, что можно сформировать набор из n ортогональных двоичных последовательностей, каждая длиной n, для n, являющегося любой степенью 2 (см. Digital Communications with Space Applications, S.W. Golomb et el. , Prentice-Hall, Inc., p.45-64). В принципе также известны наборы ортогональных бинарных последовательностей для большинства длин, которые кратны четырем и меньше двухсот. Один класс таких последовательностей, которые легко генерировать, называется функцией Уолша, известной так же, как матрица Адамара.

Функция Уолша n-го порядка может быть определена рекурсивно следующим образом:

$$W(n) = \begin{vmatrix} W(n/2), W(n/2) \\ W(n/2), W'(n/2) \end{vmatrix}$$

где W' обозначает логическое дополнение W, а W(1)= |0|.

Таким образом,

$$W(2) = \begin{vmatrix} 0,0 \\ 0,1 \end{vmatrix}$$

$$W(4) = \begin{vmatrix} 0,0,0,0 \\ 0,1,0,1 \\ 0,0,1,1 \\ 0,1,1,0 \end{vmatrix}$$

и

$$W(8) = \begin{vmatrix} 0,0,0,0,0,0,0,0 \\ 0,1,0,1,0,1,0,1 \\ 0,0,1,1,0,0,1,1 \\ 0,1,1,0,0,1,1,0 \\ 0,0,0,0,1,1,1,1 \\ 0,1,0,1,0,1,0,1 \\ 0,0,1,1,1,1,0,0 \\ 0,1,1,0,1,0,0,1 \end{vmatrix}$$

Одной из строк матрицы функции Уолша является символ, последовательность или код Уолша. Матрица функции Уолша n-го порядка содержит n последовательностей, каждая из которых имеет длину n элементов Уолша. Каждый код Уолша имеет соответствующий индекс Уолша, где индекс Уолша относится к числу /от 1 до n/, соответствующему строке, в которой находится код Уолша. Например, для приведенной выше матрицы функции Уолша с n=8 все нулевые строки соответствуют индексу Уолша 1, а код Уолша 0,0,0,0,1,1,1,1, соответствует индексу Уолша 5.

Матрица функции Уолша n-го порядка (а также все другие ортогональные функции длиной n) обладают свойством, состоящим в том, что на интервале n бит взаимная корреляция между всеми несхожими последовательностями внутри набора равна нулю. Это вытекает из того, что каждая последовательность отличается от любой другой последовательности ровно половиной своих бит. Следует также отметить, что всегда существует одна последовательность, содержащая все нули, и что все другие последовательности содержат половину единиц и половину нулей. Символ Уолша, который состоит из всех логических нулей вместо половины из нулей и половины из единиц, называется нулевым символом Уолша.

В канале обратной линии связи от удаленного устройства радиосвязи к базовой станции отсутствует пилот-сигнал, обеспечивающий привязку по фазе. Таким образом, имеется необходимость в способе, обеспечивающем высококачественную связь по каналу с замиранием, имеющему низкое отношение Eb/No/ отношение энергии на один бит к плотности мощности шума/. Модуляция функции Уолша в обратной линии связи является простым способом получения 64-ричной модуляции с когерентностью для набора из шести кодовых символов, отображенных в 64 кода Уолша. Характеристики наземного канала таковы, что частота изменения фазы относительно низка. Следовательно, путем выбора длины кода Уолша, короткого по сравнению с частотой изменения фазы в канале, возможна когерентная демодуляция на длине одного кода Уолша.

В канале обратной линии связи код Уолша определяется информацией, передаваемой из удаленного устройства радиосвязи. Например, трехбитовый информационный символ может быть отображен в приведенные выше восемь последовательностей W(8). "Обратное отображение" закодированных символов Уолша в оценку исходных информационных символов может быть выполнено в приемнике с помощью быстрого преобразования Адамара /БПА/ Предпочтительное "обратное отображение", или процесс селекции, дает "мягкое" /программируемое/решение, которое может быть передано в декодер для декодирования по критерию максимального правдоподобия.

Процедура БПА используется для выполнения процесса "обратного отображения". Процедура БПА коррелирует принятую последовательность с каждой из возможных последовательностей Уолша. Для выбора наиболее вероятного значения корреляции, которое масштабируется и принимается в качестве "мягкого" решения, используется схема селекции.

Приемник сигнала с расширенным спектром с разнесенным приемом или многоканальный приемник /"Rake" - приемник/ содержит множество приемников данных для смягчения эффекта замирания. Обычно каждый приемник данных предназначается для демодулирования сигнала, пришедшего по своей, отличной от других трассе распространения, либо с использованием многоэлементных антенн, либо с использованием свойств многолучевого распространения канала. При демодуляции сигналов, модулированных в соответствии со схемой передачи ортогональных сигналов, каждый приемник данных коррелирует принимаемый сигнал с каждым из возможных значений отображения, используя процедуру БПА, БПА каждого приемника данных

объединяются, и затем схема селекции выбирает наиболее вероятное значение корреляции, основываясь на максимальном объединенном выходном сигнале БПА для получения демодулированного символа "мягкого" решения.

В системе, описанной в Патенте США N 5103459, сигнал вызова начинается в виде источника информации 9600 бит в секунду, который затем преобразуется кодером прямого исправления ошибок со скоростью 1/3 в выходной поток 28800 символов в секунду. Эти символы группируются по 6 для образования 4800 символов Уолша в секунду, причем каждый символ Уолша отбирает одну из шестидесяти четырех ортогональных функций Уолша длительностью по шестьдесят четыре элемента Уолша. Элементы Уолша модулируются с помощью генератора псевдошумовой последовательности, специфической для каждого пользователя. Затем данные, модулированные выделенной для каждого пользователя специфической псевдошумовой последовательностью, расщепляются на два сигнала, один из которых модулируется с помощью ПСП синфазного (I) канала, а другой модулируется с помощью ПСП квадратурного (Q) канала. Как I-канальная, так и Q-канальная модуляция дает четыре псевдошумовых элемента на один элемент Уолша с частотой псевдослучайного кода расширения спектра 1.2288 МГц. I- и Q-модулированные данные представляют собой квадратурную фазовую модуляцию со сдвигом, объединенную для передачи.

В сотовой системе МДКР, описанной в вышеупомянутом Патенте США N 4901307, каждая базовая станция обеспечивает в ограниченной территориальной зоне и связывает удаленные устройства радиосвязи в зоне обслуживания с помощью коммутатора сотовой системы с коммутируемой телефонной сетью общего пользования. Когда удаленное устройство радиосвязи приближается к зоне обслуживания другой базовой станции, маршрутизация вызова этого пользователя передается новой базовой станции. Канал передачи сигнала от базовой станции к удаленному устройству радиосвязи называется прямой линией связи, а канал передачи сигнала от удаленного устройства радиосвязи на базовую станцию называется обратной линией связи.

Как было описано выше, интервал элемента псевдошумового кода определяет минимальное разнесение, которое должны иметь две трассы распространения, чтобы их можно было объединить. Прежде чем демодулировать разные трассы распространения сигналов, необходимо сначала определить относительные времена прихода /или сдвиги/ сигналов для разных трасс распространения в принимаемом сигнале. Модем канального элемента выполняет эту функцию посредством "поиска" в последовательности возможных сдвигов для трасс распространения и изменения энергии, принимаемой при каждом таком возможном сдвиге. Если энергия, связанная с возможным сдвигом, превышает некоторый порог, то такому сдвигу может быть присвоен элемент демодуляции сигнала. Затем сигнал, соответствующий этому сдвигу для трасс расширения, может быть просуммирован с составляющими от других элементов демодуляции для соответствующих сдвигов. Способ и устройство определения элементов демодуляции на основе оценки уровней энергии элементов демодуляции поисковой системы раскрыты в заявке на Патент США N 08/144902 от 28 октября 1993 года, переуступленной правопреемнику настоящего изобретения. Такой приемник с разнесением, или многоканальный /RAKE/ приемник, обеспечивает надежную цифровую связь, поскольку замирание должно иметь место для всех трасс одновременно, чтобы параметры суммарного сигнала ухудшились.

На фиг. 1 в качестве примера показан набор сигналов, поступающих на базовую станцию от одного удаленного устройства радиосвязи. Вертикальная ось представляет мощность в децибелах (дБ). На горизонтальной оси указана задержка времени прихода сигнала вследствие задержек многолучевого распространения. Ось, перпендикулярная плоскости страницы (не показана), представляет сегмент времени. Каждый пик сигнала в плоскости страницы соответствует одному и тому же моменту времени, хотя передача осуществлялась удаленным устройством радиосвязи в разные моменты времени. На общей плоскости пики, лежащие правее, соответствуют сигналу, переданному удаленным устройством радиосвязи раньше, чем сигналы, соответствующие пикам, лежащим левее. Например, самый левый пик 2 соответствует самому последнему переданному сигналу. Каждый пик сигнала 2-7 соответствует прохождению по разной трассе и, следовательно, имеет разное время задержки и разную амплитудную характеристику. Шесть различных сигнальных пиков, показанных в виде пиков 2-7, характеризуют собой среду с существенным многолучевым распространением. Обычная городская среда дает меньше пригодных для использования трасс распространения. Уровень собственных шумов системы представлен пиками и провалами, имеющими более низкие уровни энергии. Задачей поискового элемента является определение задержки, измеряемой по горизонтальной оси сигнальных пиков 2-7 для распределения потенциальных элементов демодуляции. Задачей элемента демодуляции является демодуляция набора пиков многолучевого распространения для суммирования их в единый выходной сигнал. Также задачей элементов демодуляции, распределенных по пикам многолучевого распространения, является слежение за пиком, так как он может сдвигаться во времени.

Также можно считать, что по горизонтальной оси отложены единицы сдвига элементов псевдошумового сигнала. В любой данный момент базовая станция принимает множество разных сигналов от одного удаленного устройства радиосвязи, каждый из которых распространялся по своей трассе и может иметь отличную от других задержку. Сигнал от удаленного устройства радиосвязи модулируется с помощью ПСП. Также на базовой станции генерируется копия ПСП. Каждый сигнал многолучевого распространения демодулируется на базовой станции отдельно с помощью кода ПСП, синхронизированного индивидуально. Можно

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считать, что координаты горизонтальной оси соответствуют сдвигу кода ПСП, который будет использован для демодуляции сигнала с этой координатой.

Заметим, что каждый пик многолучевого распространения изменяется по амплитуде в функции времени, как это показано в виде неровного гребня каждого пика многолучевого распространения. На показанном ограниченном временном отрезке нет больших изменений в пиках многолучевого распространения. В более широком временном диапазоне пики многолучевого распространения исчезают и со временем создаются новые траектории. Пики также могут смещаться в сторону более ранних или более поздних сдвигов в результате изменения длины траектории при движении удаленного устройства радиосвязи в зоне действия базовой станции. Каждый элемент демодуляции отслеживает наибольшие изменения выделенного ему сигнала. Задачей процесса поиска является формирование описания текущей среды многолучевого распространения, воспринимаемой базовой станцией.

В обычной радиотелефонной системе связи в передатчике удаленного устройства радиосвязи может быть использована система вокодирования, которая кодирует речевую информацию в формате переменной скорости. Например, скорость передачи данных может снижаться из-за пауз в речи. Пониженная скорость передачи данных уменьшает уровень перекрестных помех для других пользователей, вызываемых передачей от удаленных устройств радиосвязи. В приемнике или в каком-то ином устройстве, связанным с приемником, используется система вокодирования для восстановления речевой информации. Вдобавок к речевой информации удаленным модулем может передаваться либо только неречевая информация, либо их сочетание.

Вокодер, подходящий для использования в такой среде, описывается в совместно поданной заявке на Патент США N 08/363170 на "Вокодер переменной скорости" от 23 декабря 1994 года, переуступленной правопреемнику настоящего изобретения. Этот вокодер из цифровых выборок речевой информации создает кодированные данные с четырьмя различными скоростями, например, примерно 8000 бит/с, 4000 бит/с, 2000 бит/с и 1000 бит/с на основе речевой активности в течение цикла длиной 20 мс. Каждый блок данных вокодера форматируется с использованием вспомогательных битов в виде кадров данных со скоростями 9600 бит/с, 4800 бит/с, 2400 бит/с и 1200 бит/с. Кадр данных максимальной скорости 9600 бит/с называется кадром полной скорости; кадр данных со скоростью 4800 бит/с называется кадром половинной скорости; кадр данных со скоростью 2400 бит/с называется кадром одной четвертой скорости и кадр данных со скоростью 1200 бит/с называется кадром одной восьмой скорости. Ни в процессе кодирования, ни в процессе форматирования кадров информация о скорости не включается в данные. Если удаленное устройство радиосвязи передает данные со скоростью, меньшей, чем полная скорость, то рабочий цикл сигнала, передаваемого удаленными устройствами радиосвязи, будет такой же, как скорость передачи данных. Например, сигнал с одной четвертой скорости от удаленного устройства радиосвязи передается только одну четвертую часть времени.

Удаленное устройство радиосвязи включает в себя рандомизатор пакетов данных. Рандомизатор пакетов данных определяет, в течение каких интервалов времени удаленное устройство радиосвязи ведет передачу и в течение каких интервалов времени он не ведет передачу при условии заданной скорости передачи данных, конкретный индентификационный номер удаленного устройства радиосвязи и время суток. При работе со скоростью, меньшей полной скорости, рандомизатор пакетов данных в составе удаленного устройства радиосвязи распределяет псевдослучайным образом интервалы активного времени внутри пакета передачи. Соответствующий рандомизатор пакетов данных включается также и в состав базовой станции, так что базовая станция может воссоздать псевдослучайное распределение на основе времени суток и конкретного идентификационного номера удаленного устройства радиосвязи, но базовая станция не знает априори скорость передачи данных передаваемого сигнала.

Интервалы времени при одной восьмой скорости определяют так называемую "учитываемую" группу временных интервалов. Удаленное устройство радиосвязи, работающее с одной четвертой скорости, ведет передачу в течение временных интервалов "учитываемой" группы и еще одного набора распределенных псевдослучайным образом выбранных интервалов. Удаленное устройство радиосвязи, работающее с половинной скоростью, ведет передачу во время временных интервалов одной четвертой скорости и другого набора распределенных псевдослучайным образом интервалов. Удаленное устройство радиосвязи, работающее с полной скоростью, ведет передачу непрерывно. Таким путем, независимо от скорости передачи данных передаваемого сигнала, каждый временной интервал, соответствующий "учитываемой" группе, однозначно определяет интервал времени, когда соответствующее удаленное устройство радиосвязи передает сигнал. Дополнительные подробности, касающиеся рандомизатора пакета данных, описываются в совместно поданной заявке на Патент США N 08/291647 на "Рандомизатор пакетов данных" от 16 августа 1994 года, переуступленной правопреемнику настоящего изобретения.

Чтобы сэкономить системные ресурсы для активных данных при передаче речи, удаленное устройство радиосвязи не передает информацию о скорости передачи данных для каждого блока данных. Следовательно, приемник должен определить скорость, при которой данные кодировались и передавались на основе передаваемого сигнала, так чтобы вокодер, связанный с приемником, мог правильно восстановить речевую информацию. Способ определения скорости, при которой кодировались пакетные данные, без получения информации о скорости от приемника, раскрываются в совместно поданной заявке на патент США N 08/233570 на "Способ и устройство для

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определения скорости передачи данных с переменной скоростью в приемнике системы связи" от 26 апреля 1994 года, переуступленной правопреемнику настоящего изобретения. Способ определения скорости данных, раскрытый в вышеупомянутой заявке, реализуется после того, как был принят и демодулирован сигнал, вследствие чего информация о скорости процесса поиска отсутствует.

На базовой станции из ансамбля принимаемых сигналов вызовов должен быть идентифицирован каждый отдельный сигнал удаленного устройства радиосвязи. Система и способ демодуляции сигнала удаленного устройства радиосвязи, принимаемого на базовой станции, описаны, например, в Патенте США N 5103459. На фиг. 2 показана блок-схема оборудования базовой станции, описанного в Патенте США N 5103459, для демодуляции сигнала удаленного устройства радиосвязи, передаваемого по обратной линии связи.

Известная типовая базовая станция содержит многоэлементное независимое поисковое устройство и элементы демодуляции. Поисковое устройство и элементы демодуляции управляются микропроцессором. В рассматриваемом в качестве примера варианте для поддержания высокой пропускной способности системы ни одно удаленное устройство радиосвязи в системе не передает пилот-сигнал. Отсутствие пилот-сигнала в обратной линии связи увеличивает время, необходимое для анализа всех возможных временных сдвигов, с которыми может быть принят сигнал от удаленного устройства радиосвязи. Обычно пилот-сигнал передается с уровнем мощности, более высоким, чем сигналы трафика, что повышает отношение сигнал-шум принимаемого пилот-сигнала по сравнению с принимаемыми канальными сигналами трафика. В противоположность этому в идеале каждое удаленное устройство радиосвязи передает сигнал обратной линии связи, который поступает с уровнем мощности, равным уровню мощности, принимаемому от любого другого удаленного устройства радиосвязи, и следовательно, имеющий низкое отношение сигнал-шум. Кроме того, канал пилот-сигнала передает известную последовательность данных. Без пилот-сигнала в процессе поиска необходимо проверять все варианты, по которым могли быть переданы данные.

На фиг. 2 в качестве примера показан вариант известной базовой станции. Базовая станция на фиг.2 имеет одну или более антенн 12, принимающих сигналы обратных линий связи удаленных устройств радиосвязи 14. Обычно зона действия городской базовой станции разделена на три субзоны, называемые секторами. При двух антеннах на один сектор обычная базовая станция имеет всего шесть приемных антенн. Принимаемые сигналы преобразуются с понижением частоты до полосы частот модулирующих сигналов аналоговым приемником 16, который разбивает сигнал на I и Q каналы и посылает эти цифровые значения по сигнальным шинам 18 в модем канального элемента 20. Обычная базовая станция содержит множество модемов канальных элементов, таких как модем канального элемента 20 /на фиг. 2 не показаны/. Каждый модем канального элемента 20 поддерживает одного пользователя. В предпочтительном варианте модем канального элемента 20 содержит четыре элемента демодуляции 22 и восемь поисковых устройств 26. Микропроцессор 34 управляет работой элементов демодуляции 22 и поисковых устройств 26. Псевдошумовой код пользователя в каждом элементе демодуляции 22 и поисковом устройстве 26 настраивается на псевдошумовой код удаленного устройства радиосвязи, выделенный для этого модема канального элемента 20. Микропроцессор 34 пошагово просматривает поисковые устройства 26, используя набор сдвигов, называемый поисковым окном, которое потенциально содержит пики сигнала многолучевого распространения, подходящие для распределения их элементам демодуляции 22. Для каждого сдвига поисковое устройство 26 сообщает микропроцессору 34 уровень энергии, который оно обнаружило в этом сдвиге. Затем микропроцессор 34 присваивает элементы демодуляции 22 трассам распространения, идентифицированным поисковыми устройствами 26. Как только один из элементов демодуляции 22 зафиксировал сигнал на распределенном ему сдвиге, он затем следит за этой трассой сам без контроля со стороны микропроцессора 34, пока на ней не возникнет замирание или пока этот элемент не будет распределен микропроцессором 34 новой трассе распространения сигнала.

В системе по фиг.2 каждый элемент демодуляции 22 и поисковое устройство 26 содержит один процессор БПА 52, способный выполнять одно преобразование БПА в течение интервала времени, равного интервалу символа Уолша. Процессор БПА функционирует в "реальном времени" в том смысле, что одно значение интервала символа Уолша вводится и значение одного символа выводится из процессора БПА. Следовательно, для обеспечения быстрого процесса поиска необходимо использовать больше, чем одно поисковое устройство 26. Каждое поисковое устройство 26 подает на микропроцессор 34 результаты выполненного поиска. Микропроцессор 34 сводит эти результаты в таблицы для использования при присвоении элементов демодуляции 22 поступающим сигналам.

На фиг. 2 показана внутренняя структура только одного элемента демодуляции 22, но понятно, что она применима также и для поисковых устройств 26. Каждый элемент демодуляции 22 или поисковое устройство 26 модема канального элемента имеет соответствующие генераторы 36,38 I- и Q-ПСП и генератор 40 специфически для каждого пользователя ПСП, который используется для выбора конкретного удаленного устройства радиосвязи. Выходной сигнал специфической для каждого пользователя ПСП 40 подвергается операции "исключающее ИЛИ" с помощью соответствующих логических элементов 42 и 44 вместе с выходными сигналами генераторов 36 и 38 I- и Q-ПСП для получения ПСП-1' и ПСП-Q', которые подаются на устройство сжатия 46. Опорные

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синхронизирующие сигналы генераторов 36, 38, 40 настраиваются на сдвиг распределенного сигнала, так что устройство сжатия 46 коррелирует принимаемые антенной I- и Q-канальные выборки с ПСП-I' и ПСП-Q', согласованной с распределенным сдвигом сигнала. Четыре выхода устройств сжатия, соответствующие четырем псевдошумовым элементам на элемент Уолша, суммируются, образуя один элемент Уолша с помощью сумматоров 48 и 50. Затем накопленный элемент Уолша подается в процессор БПА. Когда получены 64 элемента, соответствующие одному символу Уолша, процессор БПА 52 коррелирует набор из 64 элементов Уолша с каждым из возможных 64 переданных символов Уолша и выдает 64 элементную матрицу данных "мягкого" решения. Затем выход процессора БПА 52 суммируется с выходами других расширенных элементов демодуляции с помощью сумматора 28. Выход сумматора 28 представляет собой демодулированный символ "мягкого" решения, взвешенный посредством доверительного уровня, который точно идентифицирует исходно переданный символ Уолша. Затем данные "мягкого" решения подаются в декодер прямого исправления ошибок 29 для дальнейшей обработки, чтобы восстановить исходный сигнал вызова. Затем этот сигнал вызова посылается через цифровую линию связи, такую как линия связи T1 или E1, которая направляет вызов в коммутируемую телефонную сеть общего пользования 32.

Как и каждый элемент демодуляции 22, каждое поисковое устройство 26 содержит тракт данных демодуляции процессором БПА, способным выполнять одно преобразование БПА в течение интервала времени, равного интервалу символа Уолша. Поисковое устройство 26 отличается от элемента демодуляции только тем, как используется его выходной сигнал и тем, что оно не обеспечивает временное слежение. Для каждого обрабатываемого сдвига каждое поисковое устройство 26 определяет энергию корреляции на этом сдвиге путем сжатия антенных выборок, накопления их в элементы Уолша, выполнения процедуры БПА и суммирования максимальной энергии выхода для каждого символа Уолша, на котором поисковое устройство задерживается при сдвиге. Окончательная сумма сообщается обратно микропроцессору 34. Обычно каждое поисковое устройство 26 в группе с другими по очереди опрашивается через поисковое окно микропроцессором 34, причем каждое из них отделено от соседнего на половину элемента псевдошумового кода. Таким образом, на каждую максимально возможную ошибку сдвига на четверть элемента приходится достаточно энергии корреляции для обеспечения того, чтобы трасса не была пропущена из-за того, что поисковое устройство не установило связь с точным сдвигом для данной трассы распространения. После последовательного просмотра поисковых устройств 26 посредством окна поиска микропроцессор 34 оценивает сообщаемые результаты и ищет трассы с наиболее мощным сигналом для распределения элементов демодуляции, как это описано в вышеупомянутой заявке на патент США N 08/144902.

Среда многолучевого распространения постоянно изменяется, так как удаленное устройство радиосвязи и другие отражающие объекты перемещаются в зоне действия базовой станции. Количество поисков, которые должны быть выполнены, определяется необходимостью достаточно быстро определить условие многолучевого распространения, так чтобы элементы модуляции могли эффективно использовать верно найденные трассы распространения сигналов. С другой стороны, необходимое количество элементов демодуляции является функцией количества упомянутых трасс, найденных для использования в любой момент времени. Для удовлетворения этих требований система по фиг.2 имеет два поисковых устройства 26 и один элемент демодуляции 22 для каждой из четырех используемых интегральных схем /ИС/ демодуляции, всего четыре элемента демодуляции и восемь поисковых устройств на один модем канального элемента. Каждый из этих двенадцати обрабатывающих элементов содержит полный тракт демодуляции данных, включающий процессор БПА, который занимает большую часть дорогостоящей площади интегральной схемы. Вдобавок к четырем ИС демодулятора, модем канального элемента также имеет ИС модулятора и ИС декодера прямого исправления ошибок, всего 6 ИС. Для управления и координации элементов демодуляции и поисковых устройств требуется мощный и дорогой микропроцессор. Как показано на фиг. 2, эти схемы полностью независимы и требуют непосредственного управления со стороны микропроцессора 34 для отслеживания корректных сдвигов и обработки выходных данных БПА. Микропроцессор 34 получает прерывание на каждый символ Уолша, чтобы обработать выходные данные БПА. Такая скорость прерывания сама по себе делает необходимым использование мощного микропроцессора.

Можно было бы обеспечить преимущество, если бы шесть ИС, необходимых для модема, удалось свести к одной ИС, в меньшей степени нуждающейся в поддержке микропроцессора, что уменьшило бы стоимость ИС и стоимость изготовления модема на уровне плат и позволило перейти к использованию более дешевого микропроцессора /или, как вариант, одного мощного микропроцессора, поддерживающего сразу несколько модемов канальных элементов/. Недостаточно просто сократить размеры в процессе производства ИС и заменить шесть микросхем на одну. Основная архитектура поискового устройства должна быть разработана заново для высокоэффективного модема на одной микросхеме. Исходя из вышесказанного, должно быть ясно, что имеется потребность в устройстве для приема и обработки сигнала, которое может демодулировать сигнал вызова с расширенным спектром при низкой стоимости и более эффективной архитектуре.

В настоящем изобретении можно использовать набор описанных выше поисковых устройств, работающих в реальном времени, или один поисковый процессор в интегральном исполнении, который может быстро оценивать большое

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число сдвигов, потенциально содержащихся в многолучевом принимаемом сигнале вызова.

Настоящее изобретение представляет собой способ поиска сигнала многолучевого распространения, который передается с неизвестной изменяемой скоростью и использует управление уровнем мощности.

Сущность изобретения

Настоящее изобретение представляет собой способ поиска сигнала с многолучевым распространением, который передается с неизвестной изменяемой скоростью и подвергается регулированию уровня мощности. Способ поиска является линейным, в том смысле, что не предпринимается попытка синхронизации процесса поиска с известным временем для содержания данных. Процесс поиска синхронизируется с границами групп управления мощности, чтобы можно было получить точные оценки мощности.

Краткое описание чертежей

Признаки, задачи и преимущества настоящего изобретения станут более очевидными из последующего подробного описания, вместе с чертежами, на которых одинаковые ссылочные символы идентифицируют соответствующие элементы и на которых показано следующее:

фиг. 1 - иллюстрация состояния сигнала в условиях существенного многолучевого распространения;

фиг.2 - блок-схема известной системы демодуляции сети связи;

фиг.3 - пример телекоммуникационной системы с МДКР, выполненной согласно настоящему изобретению;

фиг. 4 - блок-схема модема канального элемента, выполненного согласно настоящему изобретению;

фиг.5 - блок-схема процессора поиска;

фиг.6 - иллюстрация циклического характера буфера антенных выборок, использующего первый сдвиг;

фиг. 7 - иллюстрация циклического характера буфера антенных выборок для второго накопления при первом сдвиге по фиг. 6;

фиг. 8 - иллюстрация циклического характера буфера антенных выборок для второго сдвига;

фиг. 9 - график, показывающий, каким образом устройство поиска обрабатывает входной сигнал приемника в функции времени;

фиг.10 - блок-схема входного каскада устройства поиска;

фиг.11 - блок-схема устройства сжатия устройства поиска;

фиг.12 - блок-схема процессора результата устройства поиска;

фиг.13 - блок-схема логического устройства упорядочения устройства поиска;

фиг.14 - временная диаграмма, показывающая последовательность обработки, изображенной на фиг.5, и соответствующее состояние некоторых элементов логического устройства управления, представленных на фиг.13; и

фиг.15 - альтернативная блок-схема процессора поиска.

Описание предпочтительного варианта осуществления изобретения

В последующем описании способа и системы для обработки телефонных вызовов в цифровой радиотелефонной системе даются различные ссылки на процессы и этапы, которые выполняются для достижения желаемого результата. Следует понимать, что такие ссылки относятся не к действиям или умственным операциям, осуществляемым человеком, а к работе, видоизменению и преобразованию различных систем, включая особенно такие системы, в которых обрабатываются электрические и электромагнитные сигналы и заряды, оптические сигналы или их комбинации. В основе таких систем лежит использование различных информационных запоминающих устройств, часто называемых "памятью", которые запоминают информацию посредством размещения и упорядочения атомных или субатомных заряженных частиц на носителе жесткого диска или в кремнии, арсениде галлия или другой полупроводниковой среды, являющейся основой интегральных схем, а также использование различных устройств обработки информации, часто называемых микропроцессорами, которые изменяют свои параметры и состояние, реагируя на указанные электрические и электромагнитные сигналы и заряды. Также предусмотрена возможность использования памяти и микропроцессоров, которые запоминают и обрабатывают энергию излучения или частицы, имеющие специальные оптические характеристики, или их комбинации, и их применение согласуется с процессом функционирования описываемого изобретения.

Настоящее изобретение может быть реализовано в самых разных системах передачи данных, а в предпочтительном варианте, показанном на фиг. 3, изобретение реализуется в системе 100 для передачи речи и данных, в которой системный контроллер и коммутатор 102 выполняет функции интерфейса и управления, позволяя установить связь с удаленными устройствами радиосвязи 104 через базовые станции 106. Блок контроллера и коммутатора /БКК/ 102 управляет маршрутизацией вызовов между коммутируемой телефонной сетью общего пользования /КТСОП/ 108 и базовыми станциями 106 для передачи на удаленные устройства радиосвязи 104 и от них.

На фиг. 4 показаны модемы канальных элементов 110 A - 110 N и другие элементы инфраструктуры базовой станции, работающие в соответствии со способами МДКР, и форматами данных, описанными в вышеуказанных патентах. Множество антенн 112 подают принимаемый сигнал обратной линии связи 114 в аналоговый приемопередатчик 116. Аналоговый приемопередатчик 116 осуществляет преобразование сигнала обратной линии связи 114 с понижением частоты до полосы частот модулирующих сигналов и дискретизирует сигнал этой полосы частот при восьмикратной частоте псевдошумовых элементов принимаемого сигнала МДКР, как было определено выше. Аналоговый приемопередатчик 116 подает цифровые антенные выборки в модемы канальных элементов 110 A - 110 N посредством сигнала объединительной платы радиоприемника базовой станции. Каждый модем канального элемента 110 A - 110 N может быть присвоен

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одному удаленному устройству радиосвязи, имеющему активную связь, установленную с базовой станцией. Все модемы канальных элементов 110 А - 110 N практически идентичны по структуре.

Если активному вызову присвоен модем канального элемента 110 А, то входной каскад демодулятора 122 и интегральный поисковый процессор 128 выделяют сигнал от соответствующего удаленного устройства радиосвязи из множества сигналов вызовов, содержащихся в сигнале обратного трактата 114, путем использования ПСП, описанных в вышеуказанных патентах и патентных заявках. Модем канального элемента 110 А включает в себя однокристальный поисковый процессор 128 для идентификации сигналов многолучевого распространения, которые могут быть использованы входным каскадом демодулятора 122. В предпочтительном варианте процессор БПА с квантованием времени 120 обслуживает как интегральный поисковый процессор 128, так и входной каскад демодулятора 122. В отличие от совместно используемых процессора БПА 120 и блока определения относительного максимума 160, интегральный поисковый процессор 128 является автономным, самоуправляемым и независимым устройством. Поисковая архитектура детально описывается в одновременно рассматриваемой заявке на патент США N 08/316177 на "Процессор сигналов многолучевого распространения для системы связи множественного доступа с расширенным спектром" от 30 сентября 1994 года, переуступленной правопреемнику настоящего изобретения.

Процессор БПА 120 является ключевым устройством в процессе демодуляции. В предпочтительном варианте процессор БПА 120 соотносит принимаемые значения символов Уолша с каждым из возможных значений символов Уолша, которые могли быть переданы удаленным устройством радиосвязи. Процессор БПА 120 выдает энергию корреляции, соответствующую каждому из возможных символов Уолша, причем более высокий уровень энергии корреляции соответствует более высокой вероятности того, что удаленным устройством радиосвязи был передан символ, соответствующий этому индексу Уолша. Затем блок определения максимума 160 определяет самый большой из 64 выходных уровней энергии преобразования. Затем максимальная энергия корреляции и соответствующий индекс Уолша из блока определения максимума 160 и каждый из 64 выходных сигналов энергии корреляции из процессора БПА 120 подаются в конвейерный процессор демодулятора 126 для дальнейшей обработки. Максимальная энергия корреляции и соответствующий индекс Уолша из блока определения максимума 160 возвращаются обратно в интегральный поисковый процессор 128.

Конвейерный процессор демодулятора 126 синхронизирует и суммирует символьные данные, получаемые на различных сдвигах, в единый поток демодулированных символов "мягкого" /программируемого/ решения. Вдобавок конвейерный процессор демодулятора 126 вычисляет уровень мощности принимаемого сигнала. Исходя из принимаемого уровня мощности формируется указание, предписывающее удаленному устройству радиосвязи повысить или понизить мощность передачи этого модуля. Команда на управление мощностью проходит через модулятор 140, который добавляет эту команду в сигнал, передаваемый базовой станцией для приема данным удаленным устройством радиосвязи. Этот контур управления мощностью функционирует согласно способу, описанному в вышеуказанном Патенте США N 5056109.

Поток символов "мягкого" решения выводится из конвейерного процессора демодулятора 126 в обратный перемежитель/декодер с прямым исправлением ошибок 130 через микропроцессорный шинный интерфейс 134. Затем данные направляются через обратный цифровой канал 121 в БКК 102, который осуществляет соединение по вы зову с КТС ОП 108.

Канал обработки данных прямой линии связи реализует функции, обратные по сравнению с обратной линией связи. Сигнал подается из КТС ОП 108 через БКК 102 в обратный цифровой канал 121. Обратный цифровой канал 121 подает сигнал на вход кодера/перемежителя 138 через микропроцессор канального элемента 136. После кодирования и перемежения данных кодер/перемежитель 138 подает данные на модулятор 140, где они модулируются, как было описано в вышеуказанных патентах. Выходной сигнал 146 модулятора 140 подается в сумматор передатчика 142, где он добавляется к выходным сигналам других модемов канальных элементов 110 В - 110 N, прежде чем они подвергнутся преобразованию с повышением частоты относительно полосы модулирующих сигналов и усилятся в аналоговом приемопередатчике 116. Способ суммирования раскрыт в совместно поданной заявке на патент США N 08/316 156 на "Последовательный элемент соединения для суммирования множества цифровых сигналов" от 30 сентября 1994 года, переуступленной правоприемнику настоящего изобретения. Как показано в вышеуказанной патентной заявке, сумматор передатчика, соответствующий каждому из модемов канального элемента, может быть включен последовательно в топологию типа "цепочки" в случайном порядке, выдавая результирующий суммарный сигнал, который подается в аналоговый приемопередатчик 116 для трансляции.

На фиг. 5 показаны элементы, входящие в интегральный поисковый процессор 128. Ключевую роль в процессе поиска играет процессор БПА с квантованием времени 120, который, как упоминалось выше, совместно используется интегральным поисковым процессором 128 и входным каскадом демодулятора 122 /на фиг. 5 не показан/. Процессор БПА 120 может выполнять преобразования символов Уолша со скоростью в 32 раза большей, чем процессор БПА на фиг. 2. Такая способность быстрого преобразования делает возможным функционирование модема канального элемента 110 с квантованием времени.

В предпочтительном варианте процессор БПА 120 построен с использованием

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6-ступенчатой цепи типа "бабочки". Как подробно описано выше, функция Уолша n-го порядка может быть рекурсивно определена следующим образом:

$$W(n) = \begin{vmatrix} W(n/2), W(n/2) \\ W(n/2), W'(n/2) \end{vmatrix}$$

где W' обозначает логическое дополнение W, а W(1)=0.

В предпочтительном варианте последовательность Уолша генерируется при n= 6, и, следовательно, для соотнесения 64 элементов Уолша одного переданного символа Уолша с каждого из 64 возможных последовательностей Уолша используется 6-ступенчатая матрицы-"бабочки". Структура и способ функционирования процессора БПА 120 подробно раскрыты в совместно поданной заявке на патент США N 08/173 460 на "Способ и устройство для выполнения быстрого преобразования Адамара" от 22 декабря 1993 года, переуступленной правопреемнику настоящего изобретения.

Чтобы воспользоваться преимуществами процессора БПА 120, имеющего тридцатидвухкратную производительность по сравнению с прототипом, работающим в реальном времени, процессор БПА 120 должен быть обеспечен данными для обработки, вводимыми с высокой скоростью. Для удовлетворения этого требования буфер антенных выборок 172 должен быть рассчитан соответствующим образом. Запись в и считывание из буфера антенных выборок 172 осуществляется циклически.

Процесс поиска группируется на наборы поисков отдельных сдвигов. Наивысший уровень группирования - это набор поисков для антенны. Каждый набор поисков для антенны состоит из множества поисковых окон. Обычно каждое поисковое окно в наборе поисков для антенны представляет собой идентично выполняемую группу поисков, где каждое поисковое окно при антенном поиске получает данные от определенной антенны. Каждое поисковое окно выполнено из ряда групп поиска, представляющих собой набор последовательных поисковых сдвигов, выполняющихся в течение времени, эквивалентного длительности символа Уолша. Каждая такая группа поиска состоит из набора элементов группы. Каждый элемент представляет собой единичный поиск на данном сдвиге.

В начале процесса поиска микропроцессор канального элемента 136 посылает параметры, определяющие поисковое окно, которое может быть частью набора антенных поисков. Ширина поискового окна может быть указана в элементах псевдошумового кода. Количество поисковых элементов, необходимых для завершения поискового окна, изменяется в зависимости от количества элементов псевдошумового кода, определенных в поисковом окне. Количество элементов на одну группу поиска может быть определено микропроцессором канального элемента 136 или может быть зафиксировано в виде некоторой константы.

На фиг.1 в качестве примера показан набор сигналов, поступающих на базовую станцию от одного удаленного устройства радиосвязи, из которого станет более понятной взаимосвязь между поисковым окном, группой поиска и элементом группы. Вертикальная ось на фиг. 1 представляет принимаемую мощность в децибелах /дБ/. По горизонтальной оси откладывается задержка по времени прихода сигнала, возникающая из-за задержек многолучевого распространения. Ось /не показана/, идущая перпендикулярно плоскости страницы, представляет сегмент времени. Все пики сигнала в плоскости страницы пришли в одно и то же время, хотя и были переданы удаленным устройством радиосвязи в разное время.

Можно считать, что горизонтальная ось масштабируется в единицах сдвига элементов псевдошумового кода. В любой данный момент времени базовая станция принимает множество различных сигналов от одного удаленного устройства радиосвязи, каждый из которых прошел до отличающейся траектории и может иметь отличную от других задержку. Сигнал удаленного модуля модулируется с помощью ПСП. На базовой станции также генерируется копия ПШП. На базовой станции, если каждый сигнал многолучевого распространения был модулирован индивидуально, появляется необходимость синхронизации кода ПСП с каждым сигналом. Каждая из этих синхронизированных ПСП на базовой станции будет иметь задержку по отношению к эталонному нулевому сдвигу из-за вышеуказанной задержки. По горизонтальной оси может быть отложено количество псевдошумовых элементов, на которое задерживается синхронизированная ПСП по отношению к нулевому эталонному сдвигу базовой станции.

На фиг. 1 сегмент времени 10 представляет набор поискового окна из сдвигов элементов псевдошумового кода, подлежащих обработке. Сегмент времени 10 разделен на пять различных групп поиска, таких как сегмент времени 9. Каждая группа поиска. в свою очередь, состоит из ряда элементов, представляющих действительные сдвиги, являющиеся объектом поиска. Например, на фиг.1 каждая группа поиска состоит из 8 различных элементов, таких как 8.

Для обработки одного элемента группы поиска 8 необходим набор выборок во времени с данным сдвигом. Например, для обработки элемента 8 требуется сжатие набора выборок при сдвиге 8, идущего назад от плоскости страницы по оси времени. Также необходимо сжатие соответствующей ПСП. ПСП может быть определена путем записи времени поступления выборок и сдвига, который необходимо обработать. Требуемый сдвиг может быть объединен с временем прихода для определения соответствующей ПСП, подлежащей корреляционной обработке с принимаемыми выборками.

Когда элемент группы поиска сжимается, выборки с приемной антенны и ПСП принимают ряд значений во времени. Заметим, что выборки с приемной антенны одинаковы для всех сдвигов, показанных на фиг.1, а пики 2-7 показывают в качестве примера пики многолучевого распространения, которые поступают одновременно и различаются только с помощью процесса сжатия.

В предпочтительном варианте, описываемом ниже, каждый элемент группы поиска смещен во времени относительно

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предыдущего элемента на половину элемента псевдошумового кода. Это означает, что если элемент группы поиска 8 подвергался корреляционной обработке, начиная с показанной на чертеже плоскости среза по направлению вперед во времени /в плоскость страницы, как показано на чертеже/, то тогда элемент слева от указанного элемента 8 будет использовать выборки, начиная с половины элемента во времени назад от показанной плоскости среза. Это продвижение во времени позволяет каждый элемент в общей группе поиска подвергать корреляционной обработке с использованием одной и той же ПСП.

Каждое удаленное устройство радиосвязи принимает сигнал, передаваемый базовой станцией, с задержкой на некоторую величину, возникающую из-за задержки при прохождении через наземную среду. В удаленном устройстве радиосвязи также генерируется один и тот же I и Q короткий псевдошумовой код и длинный псевдошумовой код пользователя. Удаленное устройство радиосвязи генерирует опорные синхронизирующие сигналы на основе опорных синхронизирующих сигналов, которые он принимает от базовой станции. Удаленное устройство радиосвязи использует опорный синхронизирующий сигнал в качестве входного сигнала генераторов I- и Q-псевдошумового короткого и длинного кода пользователя. Информационный сигнал, принимаемый на базовой станции от удаленного устройства радиосвязи, подвергается двухсторонней /в прямом и обратном направлении/ задержке на пути между базовой станцией и удаленным устройством радиосвязи.

Следовательно, если синхронизация псевдошумового генератора, используемого в процессе поиска, подчиняется опорным синхронизирующим сигналам нулевого сдвига на базовой станции, то выходной сигнал генератора будет всегда появляться прежде, чем соответствующий сигнал будет принят от удаленного устройства радиосвязи.

В сигнале квадратурной фазовой модуляции со сдвигом /КФМнС/ данные I-канала и Q-канала сдвинуты друг относительно друга во времени на половину элемента. Следовательно, сжатие КФМнС, используемое в предпочтительном варианте, требует выборки данных с двойной частотой элементов. Также процесс поиска протекает оптимально с данными, дискретизируемыми с половинной частотой элементов. Каждый элемент в группе поиска сдвинут на половину элемента от предыдущего элемента. Разрешающая способность в половину элемента группы поиска обеспечивает то, что пиковые сигналы многолучевого распространения не пропускаются, не будучи обнаруженными. По этим причинам буфер антенных выборок 172 на фиг. 5 запоминает выбранные данные с удвоенной частотой псевдошумовых элементов.

Эквивалент одного символа Уолша считывается из буфера антенных выборок 172 для обработки одного элемента группы поиска. Для каждого последующего элемента из буфера антенных выборок 172 считывается эквивалент одного символа Уолша, сдвинутого на половину псевдошумового элемента от предыдущего элемента. Каждый элемент группы поиска сжимается с помощью той же самой ПСП, считываемой из буфера ПСП 176 устройством сжатия 178 для каждого элемента в группе поиска.

Буфер антенных выборок 172 имеет емкость в два символа Уолша, и в него и из него непрерывно и многократно производится запись и считывание на протяжении всего процесса поиска. В каждой группе поиска первым обрабатывается элемент, имеющий самый последний сдвиг во времени. Самый последний сдвиг соответствует сигналу, который пришел по самой длинной траектории от удаленного устройства радиосвязи до базовой станции. Момент, с которого поисковое устройство начинает обрабатывать группу поиска, привязывается к границам символа Уолша, относящегося к элементу этой границы, имеющему самый последний сдвиг в группе поиска. Временной строб, называемый границей сдвинутого символа Уолша, указывает на самый ранний момент, когда все требуемые выборки имеются в буфере антенных выборок 172, и процесс поиска может быть начат с первого элемента в группе поиска.

Работу буфера антенных выборок 172 легко проиллюстрировать, если учесть ее циклический характер. На фиг. 6 показана диаграмма работы буфера антенных выборок 172. На фиг. 6 окружность 400, показанная жирной линией, представляет сам буфер антенных выборок 172. Буфер антенных выборок 172 содержит ячейки памяти для двух значений данных символов Уолша. Указатель записи 406 циркулирует в буфере антенных выборок 178 в направлении, указанном в реальном времени, что означает, что указатель записи 406 поворачивается в буфере антенных выборок емкостью два символа Уолша в то время, когда эквивалент выборок двух символов Уолша поступает во входной каскад поискового устройства 174. Когда выборки записываются в буфер антенных выборок 172 в соответствии с адресом ячейки памяти, указываемом указателем записи 406, ранее записанные данные переписываются. В предпочтительном варианте буфер антенных выборок 176 содержит 1024 антенных выборки, поскольку каждый из двух символов Уолша содержит 64 элемента Уолша, а каждый элемент Уолша содержит 4 псевдошумовых элемента, каждый из которых отбирается дважды.

Процесс поиска делится на дискретные временные интервалы. В предпочтительном варианте интервал времени равен 1\32 длительности символа Уолша. Выбор 32 интервалов времени на один символ Уолша получается из имеющейся тактовой частоты и количества тактовых циклов, необходимых для выполнения БПА. Для выполнения БПА для одного символа Уолша требуется 64 тактовых цикла. В предпочтительном варианте имеется тактовый генератор, работающий с восьмикратной частотой псевдошумовых элементов, который обеспечивает необходимый уровень характеристик функционирования. Восьмикратная частота псевдошумовых элементов, умноженная на требуемые 64 тактовых импульса, эквивалентна времени, которое уходит на прием эквивалента двух

элементов Уолша /данных/. Поскольку в каждой половине буфера имеется 64 элемента Уолша, для считывания всего символа Уолша необходимо 32 интервала времени.

На фиг. 6 набор концентрических дуг вне окружности 400, показанной жирной линией, представляет операцию считывания и записи в буфере антенных выборок 172. Дуги внутри окружности используются для облегчения понимания и не соответствуют операциям считывания или записи. Каждая дуга представляет операцию считывания или записи во время одного интервала времени. Дуга, ближайшая к центру окружности, появляется во времени первой, а каждая последующая дуга представляет операцию, происходящую в более поздний интервал времени, как показано с помощью стрелки времени 414. Каждая из концентрических дуг соответствует части буфера антенных выборок 172, представленного окружностью 400, которая показана жирной линией. Если представить радиусы, выходящие из центра окружности 400, к точкам по концам каждой концентрической дуги, то часть окружности 400 между точками пересечения радиусов и окружности 400 будет представлять доступные ячейки памяти. Например, в течение показанной на фигуре операции на первом интервале времени в буфер антенных выборок 172, представленный дугой 402A, записывается 16 антенных выборок.

На фиг. 6,7 и 8 в качестве примера приняты следующие параметры поиска для проискового окна:

ширина поискового окна = 24 псевдошумовых элемента;

поисковый сдвиг = 24 псевдошумовых элемента;

количество символов для накопления = 2;

количество элементов на одну поисковую группу = 24.

На фиг. 6 также принято, что буфер антенных выборок 172 практически содержит эквивалент полного символа Уолша достоверных данных перед записью, показанной дугой 402A. Во время последующих интервалов времени происходит запись, соответствующая дуге 402B и луге 402C. В течение 32 интервалов времени, имеющихся в наличии в течение времени, эквивалентного одному символу Уолша, операции записи продолжаются от дуги 402A до самой большой дуги 402 FF, которая не показана.

32 интервала времени, представленные дугами с 402A до 402FF, соответствуют времени, используемому для завершения одной поисковой группы. Используя заданные выше параметры, поисковая группа начинает сдвиг на 24 псевдошумовых элемента от эталонного нулевого сдвига, или "реального времени", и содержит в себе 24 элемента. Сдвиг на 24 псевдошумовых элемента соответствует повороту на 16.875 градуса по окружности 400 от начала первой записи, показанной дугой 402A /вычисляемому путем деления сдвига на 24 псевдошумовых элемента на общее число элементов 256 в половине буфера антенных выборок 172 и умножения результата на 180 градусов/. Эта дуга в 16.875 градуса показана в виде дуги 412. 24 элемента поисковой группы соответствуют считываниям, показанным дугами 404A-404X, самая большая из которых не показана. Первое считывание, соответствующее дуге 404A, начинается в некоторый момент поискового сдвига после записи, соответствующей 402C, так что в наличии имеется набор смежных данных. Каждое последующее считывание, например 404B, смещается по отношению к предыдущему на один адрес памяти, соответствующий 1/2 псевдошумового элемента во времени. В течение показанной поисковой группы считывания происходят по направлению к более ранним сдвигам, показанным дугами 404A-404X, с течением времени против времени против часовой стрелки и в направлении, противоположном вращению указателя записи 406. 24 считывания, представленные дугами 404A-404X, обходят дугу, показанную под номером 418. Проведение считываний по направлению к более ранним выборкам имеет то преимущество, что обеспечивается непрерывный поиск внутри поискового окна при реализации каждой поисковой группы. Это преимущество подробно раскрывается ниже.

При каждом из считываний, соответствующих дугам от 404A до 404X, на устройство сжатия 178 подается эквивалент одного значения символа Уолша /данных/. Следовательно, считывание соответствует охвату 180 градусов на окружности 400. Заметим, что в поисковой группе, показанной на фиг. 6, последняя запись, соответствующая дуге 402FF, и последнее считывание, соответствующее дуге 404X, не включают каких-либо общих адресов памяти для обеспечения смежных достоверных данных. Однако можно предположить, что если диаграмма считывания и записей продолжалась, то они в действительности пересекались бы, и в этих условиях нельзя было бы обеспечить достоверные данные.

В большинстве случаев передачи сигналов результат эквивалента элемента данных поисковой группы, получаемый в течение временного эквивалента одного символа Уолша, недостаточен для обеспечения точной информации о расположении разнесенных сигналов. В этих случаях поисковая группа может быть многократно повторена. Результаты для элементов в последующих поисковых группах с общим сдвигом накапливаются процессором 162 результатов поиска, что подробно разъяснено далее. В этом случае заданные выше параметры поиска указывают, что количество символов для накопления на каждом сдвиге равно двум. На фиг. 7 показана поисковая группа по фиг. 6, повторяемая с одним и тем же сдвигом в течение временного эквивалента последующего символа Уолша. Заметим, что буфер антенных выборок 172 содержит эквивалент двух символов Уолша, так что данные, необходимые для обработки в течение поисковой группы, показанной на фиг. 7, были записаны для поисковой группы, показанной на фиг. 6. В этой конфигурации отстоящие друг от друга на 180 градусов ячейки памяти представляют один и тот же псевдошумовой сдвиг.

После завершения двух поисковых групп по фиг. 6 и 7 процесс поиска переходит к следующему сдвигу в поисковом окне.

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Величина продвижения равна ширине обрабатываемой поисковой группы и в данном случае составляет 12 псевдошумовых элементов. Ширина окна определяет, сколько сдвигов поисковых групп необходимо для завершения поискового окна. В данном случае для перекрытия поискового окна шириной 24 псевдошумовых элемента необходимо два разных сдвига. На фиг.8 ширина окна показана дугой 412. Второй сдвиг для этого поискового окна начинается со сдвига, следующего за последним сдвигом предыдущей поисковой группы, и продолжается вокруг точки номинального нулевого сдвига, устанавливаемого по местоположению начала первой записи, указанной дугой 430А. Здесь имеется 24 элемента внутри поисковой группы, показанной дугами 432А-432Х, причем самая большая из них не показана. 32 записи показаны дугами 430А-430FF. Таким образом, последняя запись, показанная дугой 430FF, и последнее считывание, показанное дугой 432Х, граничат друг с другом в буфере антенных выборок 172.

Поисковая группа, показанная на фиг. 8, повторяется на противоположной стороне буфера антенных выборок 172 почти так же, как поисковая группа на фиг. 6 повторяется на фиг. 7, поскольку параметры поиска предписывают, чтобы каждый символ накапливался дважды. После завершения второго накопления второй поисковой группы интегральный процессор поиска 128 готов начать работу с другим поисковым окном. Следующее поисковое окно может иметь новый сдвиг, или может определять новую антенну, или и то и другое.

На фиг. 8 местоположение границы между областью считывания и областью записи буфера отмечено меткой 436. На фиг. 6 граница отмечена меткой 410. Сигнал, указывающий точку во времени, соответствующую меткам 410 и 436, называется стробом смещенного символа Уолша, который также показывает, что появился эквивалент нового символа Уолша выборок. Когда поисковые группы внутри окна продвигаются к более ранним сдвигам, граница между областями буфера для считывания и для записи поворачивается на шаге фиксации против часовой стрелки, как показано на фиг.8. Если после завершения текущего поискового окна требуется большое изменение в обрабатываемом сдвиге, то строб сдвинутого символа Уолша может быть продвинут на большую часть окружности цикла.

Фиг. 9 дает графическое представление процесса поиска во времени. Время отображается в единицах символов Уолша по горизонтальной оси. Адреса буфера антенных выборок 172 и адреса буфера 176 ПСП показаны по вертикальной оси также в единицах символов Уолша. Поскольку буфер 172 антенных выборок имеет емкость два символа Уолша, то адресация буфера 172 антенных выборок охватывает границы четного символа Уолша, но на фиг.9 для примера показаны адреса перед их сверткой один поверх другого. Выборки записываются в буфер 172 антенных выборок по адресу, взятому непосредственно с момента, когда они были получены, так что указатель записи 181 в буфере 172 антенных выборок представляет собой прямую с наклоном сорок пять градусов. Обрабатываемый сдвиг отображается в базовый адрес в буфере 174 антенных выборок для начала считывания символа Уолша выборок для одного элемента поисковой группы. Элементы показаны на фиг. 9 в виде практически вертикальных линейных сегментов указателей считывания 192. Каждый элемент отображается в символ Уолша по высоте, обозначенной на вертикальной оси, и в 1/32 символа Уолша, указанного по горизонтальной оси.

Промежутки по вертикали между элементами поисковой группы обусловлены входным каскадом 122 демодулятора, прерывающим процесс поиска, чтобы использовать процессор БПА 120. Входной каскад 122 демодулятора функционирует в реальном времени и имеет наивысший приоритет на использование процессора БПА 120, всякий раз когда он имеет текущий или очередной набор данных для обработки. Следовательно, обычно входному каскаду 122 демодулятора предоставляется возможность использования процессора БПА 120 на границе каждого символа Уолша, соответствующей сдвигу псевдошумового сигнала, который демодулируется входным каскадом 122 демодулятора.

На фиг. 9 показаны те же самые поисковые группы, что и на фиг. 6, 7 и 8. Например, поисковая группа 194 имеет 24 элемента, каждый из которых соответствует одной из дуг 404А-404Х на фиг.6. На фиг.9 для поисковой группы 194 указатель 410 показывает строб сдвинутого символа Уолша, соответствующий аналогичному указателю на фиг. 6. Для считывания текущих выборок каждый элемент поисковой группы должен находиться ниже указателя записи 181. Наклон вниз элементов вместе с поисковой группой указывает шаги к более ранним выборкам. Поисковая группа 195 соответствует поисковой группе, показанной на фиг. 7, а поисковая группа 196 соответствует поисковой группе, показанной на фиг. 8.

В поисковом окне, определенном вышеуказанными параметрами, определены только 24 элемента на одну поисковую группу, хотя данная поисковая группа имеет 32 доступных интервала времени. Каждый элемент может быть обработан в течение одного интервала времени. Однако практически невозможно увеличить количество элементов на одну поисковую группу до 32, чтобы согласовать количество квантов времени в течение поисковой группы. Входной каскад 122 демодулятора использует некоторые из имеющихся интервалов времени процессора БПА. Имеется также временная задержка, связанная с продвижением поисковой группы, когда процедура считывания должна ожидать осуществления операции записи, чтобы заполнить буфер достоверными данными для нового сдвига. Также необходим некоторый запас для синхронизации с интервалом времени границы обработки после обнаружения строба сдвинутого символа Уолша. Все эти факторы ограничивают на практике количество элементов, которые могут быть обработаны в одной поисковой группе. В некоторых случаях количество элементов на одну поисковую группу может

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быть увеличено, например, если входной каскад 122 демодулятора имеет только один присвоенный демодулирующий элемент и, следовательно, прерывает процессор БПА 120 один раз за поисковую группу. Следовательно, в предпочтительном варианте количество элементов на поисковую группу регулируется микропроцессором канального элемента 136. В альтернативных вариантах количество элементов на одну поисковую группу может быть постоянной фиксированной величиной.

Также может иметь место значительная дополнительная задержка при коммутации антенн источников на входе в буфер выборок или изменении точки начала поискового окна или интервала между поисками. Если одна поисковая группа требует конкретный набор выборок, а следующая группа для другой антенны требует использование перекрывающейся части буфера, то следующая группа должна отсрочить обработку, пока не появится следующая граница сдвинутого символа Уолша, с момента появления которой будет иметься законченный символ Уолша выборок для нового антенного источника. На фиг. 9 поисковая группа 198 обрабатывает данные от другой антенны, по сравнению с антенной для поисковой группы 197. Горизонтальная линия 188 указывает область памяти, соответствующую входным выборкам новой антенны. Отметим, что поисковые группы 197 и 198 не используют какие-либо общие области памяти.

Для каждого интервала времени в буфер выборок должны записываться два элемента Уолша, а из буфера выборок может быть считан один полный символ Уолша. В предпочтительном варианте в течение каждого интервала времени имеет место 64 тактовых цикла. Полный элемент Уолша из выборок содержит четыре набора выборок: текущие I-канальные выборки, прежние I-канальные выборки, текущие Q-канальные выборки и прежние Q-канальные выборки. В предпочтительном варианте каждая выборка состоит из четырех бит. Следовательно, за один тактовый импульс от буфера антенных выборок 172 требуется шестьдесят четыре бита. Используя однопортовое ОЗУ, самый простой вариант буфера удваивает ширину слова до 128 бит и расщепляет буфер на два 64-битовых отрезка /64-битового слова/ в виде буферов 168,170 для независимого считывания/записи четных и нечетных элементов Уолша. Затем записи, поступающие в буфер с гораздо меньшей частотой, мультиплексируются между операциями считывания, которые осуществляют переключение между двумя банками данных на последовательных тактовых циклах.

Считывание выборок элементов Уолша из четных и нечетных буферов 168,170 элементов Уолша имеет произвольное выравнивание по отношению к физическому выравниванию слова в ОЗУ. Поэтому при первом считывании интервала времени обе половины считываются в устройство сжатия 178, образуя окно шириной в два элемента Уолша, из которого получается один элемент Уолша с выравниванием по текущему сдвигу. Для поисковых сдвигов четных элементов Уолша адрес буфера для четного и нечетного элементов Уолша для первого считывания один и тот же. Для сдвигов нечетных элементов Уолша четный адрес для первого считывания сдвигается на единицу от нечетного адреса, обеспечивая начало последующего элемента Уолша от нечетной половины буфера выборок. Дополнительные элементы Уолша, необходимые для устройства сжатия 178, могут быть поданы на него посредством считывания из одного буфера элементов Уолша. Далее последующие считывания всегда обеспечивают обновленное окно шириной в два элемента Уолша, из которого извлекается элемент Уолша, выровненный с обрабатываемым в текущий момент сдвигом.

Согласно фиг. 5, для каждого элемента в поисковой группе в процессе сжатия используется один и тот же символ Уолша ПСП из буфера 176 ПСП. Для каждого тактового цикла в интервале времени необходимы четыре пары псевдошумовых выборок I'-канала и Q'-канала. При использовании однопортового ОЗУ ширина слова удваивается и считывается из половины буфера многократного. Затем в цикле, не используемом для считывания, выполняется единичная запись в буфере 176 ПСП, требуемая для одного интервала времени.

Поскольку процесс поиска может определить поиск сдвигов псевдошумового сигнала до двух символов Уолша, задержанных относительно текущего момента, то должен запоминаться эквивалент четырех символов Уолша ПСП. В предпочтительном варианте буфер 176 ПСП представляет собой ОЗУ емкостью 128 слов на 16 бит. Требуется четыре символа Уолша, поскольку начальный сдвиг может изменяться на 2 символа Уолша, и коль скоро выбран начальный сдвиг, эквивалент одного символа Уолша ПСП необходим для корреляции, что означает, что для процесса сжатия необходимо иметь эквивалент трех символов Уолша. Поскольку одна и та же ПСП используется многократно, данные в буфере 176 ПСП не могут быть переписаны во время процесса сжатия, соответствующего одной поисковой группе. Следовательно, необходимо иметь эквивалент одного дополнительного символа Уолша в памяти для запоминания данных ПСП при их генерации.

Данные, которые записываются в буфер 176 ПСП и буфер 172 антенных выборок, обеспечиваются входным каскадом устройства поиска 174. Блок-схема входного каскада устройства поиска 74 показана на фиг. 10. Входной каскад устройства поиска 174 включает в себя генераторы I и Q-ПСП 202, 206 короткого кода и генератор ПСП пользователя 204 длинного кода. Значения, выдаваемые генераторами I- и Q-ПСП 202, 206 короткого кода и генеретор ПСП пользователя 204 длинного кода частично определяются временем суток. Каждая базовая станция имеет универсальный стандарт синхронизации, например, GPS синхронизацию для создания синхронизирующего сигнала. Каждая базовая станция также передает в эфир синхронизирующий сигнал на удаленные устройства радиосвязи. На базовой станции опорные синхронизирующие сигналы имеют

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нулевой сдвиг, поскольку они настроены на опорные сигналы всемирного времени.

Выходной сигнал генератора ПСП пользователя 204 длинного кода подвергается логической операции "исключающее ИЛИ" вместе с выходным сигналом генераторов I- и Q-ПСП 202,206 коротких кодов с помощью логических элементов XOR /"исключающее ИЛИ"/ 208 и 210 соответственно. /Такой же процесс осуществляется также и в удаленном устройстве радиосвязи, и выходной сигнал используется для модуляции сигнала, передаваемого удаленным устройством радиосвязи/. Выходной сигнал логических элементов XOR 208 и 210 запоминается в последовательно-параллельном сдвиговом регистре 212.

Последовательно-параллельный сдвиговый регистр 212 буферизирует последовательности до ширины буфера 176 ПСП. Затем выходной сигнал последовательно-параллельного сдвигового регистра 212 записывается в буфер 176 ПСП по адресу, который берется с опорной отметки времени с нулевым сдвигом. Таким образом, входной каскад устройства поиска 174 подает последовательные псевдошумовые данные в буфер 176 ПСП.

Входной каскад устройства поиска 174 подает также антенные выборки в буфер 172 антенных выборок. Выборки приема 118 отбираются с одной из множества антенн мультиплексора 216. Отобранные выборки приема подаются на ключевую схему с фиксацией состояния 218, где они прореживаются, что означает, что одна четверть выборок отбирается для использования в процессе поиска. Выборки приема 118 были отобраны с восьмикратной частотой псевдошумовых элементов аналоговым приемопередатчиком 116 /по фиг.4/. Обработка согласно алгоритму поиска предназначена для выборок, отбираемых с половинной частотой псевдошумовых элементов. Следовательно, в буфер 172 антенных выборок должна проходить только четверть принимаемых выборок.

Выходной сигнал ключевой схемы с фиксацией состояния 218 подается в последовательно-параллельный сдвиговый регистр 214, который буферизирует выборки до ширины буфера 172 антенных выборок. Затем выборки записываются в буферах четных и нечетных элементов Уолша 168,170 по адресам, которые также берутся с опорной отметки времени с нулевым сдвигом. Таким способом устройство сжатия 178 может синхронизировать данные антенных выборок с известным сдвигом по отношению к ПСП.

В соответствии с фиг. 5, в течение каждого тактового цикла в интервале времени устройство сжатия 178 отбирает элемент Уолша антенных выборок из буфера 172 антенных выборок и соответствующий набор значений ПСП из буфера 176 ПСП и выводит I- и Q-канальный элемент Уолша в процессор БПА через мультиплексор 124.

На фиг.11 показана подробная блок-схема устройства сжатия 178. Ключевая схема с фиксацией состояния 220 для четных элементов Уолша 220 и ключевая схема с фиксацией состояния 222 для нечетных элементов Уолша фиксирует данные от буфера 168 четных элементов Уолша и буфера 170 нечетных элементов Уолша соответственно. Блок мультиплексоров 224 выделяет элемент Уолша выборок, подлежащих использованию, из эквивалента двух элементов Уолша выборок, представляемых ключевыми схемами с фиксацией состояния 220 и 222 для четных и нечетных элементов Уолша. Логический блок выбора 226 мультиплексора определяет границу отобранного элемента Уолша на основе сдвига обрабатываемого элемента поисковой группы. Элемент Уолша выводится на блок схем "исключающее ИЛИ" устройства сжатия /КФМн/ 228 сигнала с квадратурной фазовой манипуляцией.

Значения ПСП из буфера 176 ПСП фиксируются ключевой схемой с фиксацией состояния 234 для ПСП. Многорегистровое циклическое сдвиговое устройство 232 сдвигает выходной сигнал ключевой схемы с фиксацией 234 на основе сдвига обрабатываемого элемента и подает ПСП в блок схем "исключающее ИЛИ" устройства сжатия 228 КФМн, где осуществляется условное инвертирование антенных выборок на основе ПСП. Затем значения, получившиеся после операции "исключающее ИЛИ", суммируются посредством дерева сумматоров 230, которое выполняет операцию суммирования в процессе сжатия КФМн и затем суммирует вместе четыре выхода сжатых элементов, образуя элемент Уолша для ввода в процессор БПА 120.

В соответствии с фиг.5, процессор БПА 120 получает от устройства сжатия 178 через мультиплексор 124 64 элемента Уолша и, используя 6-ступенчатую решетку, коррелирует эти шестьдесят четыре входные выборки с каждой из шестидесяти четырех функций Уолша в интервале времени из шестидесяти четырех тактовых циклов. Для определения сигнала корреляции с наибольшей энергией от процессора БПА 120 может быть использован детектор максимума 160. Выходной сигнал детектора максимума 160 подается на процессор результатов поиска 162, являющегося частью интегрального процессора поиска 128.

Процессор результатов поиска 162 подробно представлен на фиг. 12. Процессор 162 также работает в режиме квантования времени. Поступающий на него управляющий сигнал подвергается конвейерной задержке для согласования с задержкой в два интервала времени от начала ввода элементов Уолша в процессор БПА 120 до получения выходного сигнала максимальной энергии. Как объяснялось выше, набор параметров поискового окна может предписывать, какое количество эквивалентов символов Уолша данных было накоплено, прежде чем будут обработаны результаты выбранного сдвига. При параметрах, используемых в примерах по фиг. 6,7,8 и 9, количество символов для накопления равно 2. Процессор результатов поиска 162, наряду с другими функциями, выполняет функцию суммирования.

Когда процессор результатов поиска 162 суммирует последовательно поступающие символы Уолша, он должен запоминать накопленную сумму для каждого элемента в поисковой группе. Эти накопленные суммы запоминаются в ОЗУ накопления символов Уолша 240. Результаты по каждой поисковой

группе вводятся в сумматор 242 от декодера максимума 160 для каждого элемента. Сумматор 242 суммирует существующий результат с соответствующим промежуточным значением, имеющимся в ОЗУ накопления символов Уолша 240. При накоплении последнего символа Уолша для каждого элемента поисковой группы промежуточный результат считывается из ОЗУ накопления символов Уолша 240 и суммируется с помощью сумматора 242 с конечной энергией от этого элемента для получения последнего результата поиска для данного сдвига элемента. Затем результаты поиска сравниваются с наилучшими результатами, полученными при поиске до того момента, который будет пояснен ниже.

В вышеупомянутой одновременно рассматриваемой патентной заявке США N 08\ 144902 в предпочтительном варианте элементы демодуляции распределяются на основе лучших результатов, полученных при поиске. В данном предпочтительном варианте в регистре лучших результатов 250 запоминается восемь лучших результатов. /В других вариантах может запоминаться меньшее или большее количество результатов/. Регистр промежуточных результатов 164 запоминает пиковые значения и их соответствующий ранговый порядок. Если энергия текущего поиска превышает по меньшей мере одно из значений энергии в регистре промежуточных результатов, то логический блок управления процессором промежуточных результатов 254 отбрасывает восьмой лучший результат в регистре промежуточных результатов 164 и вставляет новый результат вместе с соответствующим рангом, сдвигом псевдошумового сигнала и антенной, соответствующей данному результату элемента поисковой группы. Все результаты с меньшим рангом понижаются на один ранг. Специалистам известно большое число способов, обеспечивающих такого рода сортировку. В рамках объема данного изобретения может быть использован любой из них.

Процессор результатов поиска 162 имеет фильтр локального пика, содержащий в основной своей части компаратор 244 и ключевую схему с фиксацией состояния предыдущего значения энергии 246. Фильтр локального пика, если он включен, запрещает обновление регистра промежуточных результатов 164, даже когда энергия, определенная в результате поиска, дает право на это, если результат поиска не представляет локальный пик многолучевого распространения. Таким путем фильтр локального пика предотвращает ввод сильного, но "размазанного" многолучевого излучения в регистр промежуточных результатов 164, что в противном случае не оставило бы места для более слабых, но отчетливых сигналов многолучевого излучения, которые могут быть более верными кандидатами для демодуляции.

Реализация фильтра локального пика достаточно проста. Значение энергии суммирования предыдущего элемента поисковой группы запоминается в ключевой схеме с фиксацией состояния для предыдущего значения энергии 246.

Сумма для текущего элемента сравнивается с хранящимся значением с помощью компаратора 244. Выходной сигнал компаратора 244 указывает, какой из двух его входных сигналов больше, и этот сигнал фиксируется в логическом блоке управления процессора результатов поиска 254. Если предыдущая выборка представляла локальный максимум, то логический блок управления процессора результатов поиска 254 сравнивает предыдущий результат энергии с данными, хранящимися в регистре промежуточных результатов 164, как описывалось выше. Если фильтр локального пика выключен микропроцессором канального элемента 136, то тогда всегда выполняется сравнение с регистром промежуточных результатов 164. Если либо первый, либо последний элемент поисковой группы на границе поискового окна имеет спад, то тогда ключевая схема с фиксацией спада настраивается так, чтобы крайнее значение границы рассматривать как пик.

Простой реализации фильтра локальных пиков способствует проведение считываний по направлению к более ранним символам внутри поисковой группы. Как показано на фиг. 6, 7, 8 и 9, внутри поисковой группы каждый элемент продвигается к сигналам, поступающим во времени раньше. Такая направленность означает, что внутри поискового окна последний элемент поисковой группы и первый элемент последующей поисковой группы являются соседними в сдвиге. Следовательно, работа фильтра локального пика не должна изменяться, и выходной сигнал компаратора 244 остается достоверным при пересечении границ поисковой группы.

В конце обработки поискового окна значения, хранящиеся в регистре промежуточных результатов 164, передаются в регистр лучших результатов 250, данные из которого считываются микропроцессором канального элемента 136. Процессор результатов поиска 162, таким образом, имеет большую рабочую нагрузку от микропроцессора канального элемента 136, обработку которой в системе по фиг. 2 требуется производить независимо по результату каждого элемента поисковой группы.

В предыдущих разделах основное внимание уделялось тракту обработки данных интегрального поискового процессора 128 и подробному описанию того, как строчные антенные выборки 118 преобразуются в итоговое описание многолучевого распространения на выходе регистра лучших результатов 250. В последующих разделах детально описывается, как управляется каждый элемент в тракте данных обработки поиска.

Блок управления поиском 166 по фиг. 5 детально представлен на фиг. 13. Как упоминалось ранее, микропроцессор канального элемента 136 определяет набор параметров поиска, включающий группу антенн для поиска, запоминаемых в буфере выбора антенн 348, начальный сдвиг, запоминаемый в буфере поискового сдвига 308, количество элементов на одну поисковую группу, запоминаемое в буфере 312 ширины группы, ширину поискового окна, запоминаемую в буфере 314 ширины поиска, количество символов Уолша для накопления,

запоминаемое в буфере 316 накопления символов Уолша, и управляющее слово, запоминаемое в буфере 346 управляющего слова.

Начальный сдвиг, запоминаемый в буфере 308 поискового сдвига, определяется с разрешением в восемь элементов. Начальный сдвиг указывает, какие выборки во входном каскаде устройства поиска 174 удаляются посредством прореживания с помощью ключевой схемы с фиксацией состояния 218 на фиг. 10. Благодаря буферу антенных выборок шириной в два символа Уолша 172 в этом варианте самое большое значение начального сдвига составляет половину элемента псевдошумового кода, что меньше двух полных символов Уолша.

До этого момента была раскрыта базовая конфигурация для выполнения поиска. В действительности имеется несколько видов заранее определенных поисков. Когда удаленное устройство радиосвязи изначально пытается получить доступ в систему, оно посылает сигнал радиомаяка, называемый преамбулой, с использованием нулевого символа Уолша. Нулевой символ Уолша - это символ Уолша, содержащий все логические нули вместо единиц и нулей, как было описано выше. При выполнении поиска преамбулы устройство поиска ищет удаленное модуль, посылающий сигнал радиомаяка с нулевым символ Уолша по каналу доступа. Результатом поиска при поиске преамбулы является энергия для нулевого символа Уолша. Когда выполняется поиск канала доступа в режиме захвата, детектор максимума 160 выводит энергию для нулевого символа Уолша независимо от выявленной максимальной выходной энергии. Управляющее слово, хранящееся в буфере 346 управляющего слова, включает в себя бит преамбулы, который указывает, когда выполняется поиск преамбулы.

Как объяснялось выше, система управления мощностью согласно предпочтительному варианту измеряет уровень сигнала, получаемого от каждого удаленного устройства радиосвязи и формирует команду управления мощностью, предписывающую удаленному устройству радиосвязи повысить или понизить мощность передачи данного удаленного устройства радиосвязи. Система управления мощностью работает с использованием символов Уолша, называемом группой управления мощностью, во время работы канала трафика. /Функционирование канала трафика следует за работой канала доступа и подразумевает работу во время активного вызова/. Все символы Уолша внутри одной группы управления мощностью передаются с использованием одной и той же команды управления мощностью в удаленном устройстве радиосвязи.

Выше также говорилось, что в предпочтительном варианте осуществления настоящего изобретения сигнал, передаваемый удаленным устройством радиосвязи, имеет изменяемую скорость /передачи данных/ во время работы канала трафика. Во время процесса поиска скорость, используемая удаленным устройством радиосвязи для передачи данных, на базовой станции не известна. При накоплении последовательных символов очень важно, чтобы передатчик во время накопления не выключался. Последовательные символы Уолша в группе управления мощностью пропускаются как группа, что означает, что 6 символов Уолша, составляющие группу управления мощностью, в предпочтительном варианте, либо все пропускаются, либо все не пропускаются.

Таким образом, если параметр поиска предписывает, что множество символов Уолша должно быть накоплено во время работы канала трафика, то процесс поиска должен выравнивать каждую поисковую группу, чтобы начинать и закончить его в рамках одной группы управления мощностью. Управляющее слово, запоминаемое в буфере 346 управляющего слова, включает в себя бит выравнивания группы управления мощностью. С помощью бита выравнивания группы управления мощностью, установленного в единицу, что указывает на поиск канала трафика, процесс поиска синхронизирует по границе следующей группы управления мощностью вместо границы следующего сдвинутого символа Уолша.

Управляющее слово, запоминаемое в буфере 346 управляющего слова, также включает в себя бит включения фильтра обнаружения пика, упоминаемого ранее в связи с фиг. 8.

Устройство поиска работает либо в непрерывном режиме, либо в одношаговом режиме в соответствии с установкой бита "непрерывный /одношаговый" управляющего слова. В одношаговом режиме после выполнения поиска интегральный поисковый процессор 128 возвращается в состояние ожидания дальнейших инструкций. В непрерывном режиме интегральный поисковый процессор 128 постоянно осуществляет поиск, и в момент, когда в микропроцессор канального элемента 136 поступает сигнал о том, что получены результаты, интегральный поисковый процессор 128 начинает следующий поиск.

Блок управления поиском 166 формирует сигналы синхронизации, используемые для управления процессом поиска, который выполняется интегральным поисковым процессором 128. Блок управления поиском 166 посылает опорные синхросигналы с нулевым сдвигом в генераторы I- и Q-ПСП 202, 206 коротких кодов и генератор ПСП пользователя 204 длинного кода и сигнал включения на ключевую схему с фиксацией состояния разряжения 218 и сигнал выбора в мультиплексор 216 во входном каскаде устройства поиска 174. Он обеспечивает адреса считывания и записи для буфера 176 ПСП и буферов четных и нечетных элементов Уолша 168 и 170. Он выдает текущий сдвиг для управления работой устройства сжатия 178. Он обеспечивает опорную синхронизацию внутри интервала времени для процессора БПА 120 и определяет, использует ли процесс поиска или процесс демодуляции процессор БПА 120, путем управления мультиплексором 124 с входом БПА. Он обеспечивает несколько версий конкретных внутренних синхронизирующих стробов с конвейерной задержкой для логического блока управления процессора результатов поиска 254 по фиг. 12, разрешая ему просуммировать результаты сдвигов для

определенного числа накопленных символов Уолша. Блок управления поиском 166 обеспечивает регистр лучших результатов 250 конвейерным сдвигом и информацией для антенн, соответствующих запомненным значениям накопленной энергии.

Согласно фиг. 13, счетчик системного времени 342 синхронизируется по опорному синхросигналу с нулевым сдвигом. В предпочтительном варианте, как было подробно описано ранее, системный тактовый генератор работает с восьмикратной частотой элементов псевдошумового кода. Имеется 256 псевдошумовых элементов в символе Уолша и 6 символов Уолша в группе управления мощностью, всего для 6x256x8=12288 системных тактовых импульсов на одну группу управления мощностью.

Таким образом в предпочтительном варианте счетчик системного времени 342 содержит четырнадцатибитный счетчик, который отсчитывает 12288 системных тактовых импульсов. Входная опорная синхронизация для генераторов I- и Q-ПСП 202, 206 коротких кодов и генератора ПСП пользователя 204 длинного кода по фиг.10 во входном каскаде устройства поиска 174 берется от счетчика системного времени 342. /Выходной сигнал генератора ПСП пользователя 204 длинных кодов также формируется на основе более длинных системных опорных синхросигналов, которые не повторяются примерно в течение 50 дней. Более длинный опорный сигнал не управляется со стороны процесса поиска и действует как заранее установленное значение. Продолжение функционирования с использованием заранее установленного значения управляется счетчиком системного времени 342/. Адреса для буфера 176 ПСП и буферов 168 и 170 четных и нечетных элементов Уолша берутся от счетчика системного времени 342. Счетчик системного времени 342 фиксируется ключевой схемой 328 в начале каждого интервала времени. Выходной сигнал ключевой схемы с фиксацией 328 отбирается через адресные мультиплексоры 330,332 и 334, которые обеспечивают адреса записи, соответствующие текущему интервалу времени, когда в эти буферы осуществляется запись в некоторый более поздний момент внутри интервала времени.

Накопитель сдвига 310 отслеживает сдвиг обрабатываемого в данный момент элемента поисковой группы. Начальный сдвиг, хранящийся в буфере поискового сдвига 308, загружается в накопитель сдвига 310 при начале каждого поискового окна.

Накопитель сдвига 310 уменьшает свое значение с каждым элементом поисковой группы. В конце каждой поисковой группы, чтобы можно было повторить эту операцию для дальнейших накоплений, количество элементов на одну поисковую группу, хранящееся в буфере 312 ширины группы, вычитается из накопителя сдвига, чтобы привязать его снова к первому сдвигу в поисковой группе. Таким способом процесс поиска снова осуществляет развертку поисковой группы для накопления другого символа Уолша. Если процесс поиска осуществил развертку для текущей поисковой группы при накоплении последнего символа Уолша, то накопитель сдвига 310 уменьшает свое значение на единицу путем выбора входа "-1" мультиплексора 304 повторной поисковой группы, который осуществляет сдвиг первого элемента в следующей поисковой группе.

Выходной сигнал накопителя сдвига 310 всегда представляет сдвиг обрабатываемого в данный момент элемента и поэтому используется для управления вводом данных в устройство сжатия 178. Выходной сигнал накопителя сдвига 310 суммируется сумматорами 336 и 338 с выходным сигналом внутренней синхронизации /по интервалам времени/ от счетчика системного времени 342 для генерирования последовательности адресов внутри интервала времени, соответствующего элементу поисковой группы. Выход сумматоров 336 и 338 отбирается через мультиплексоры 330 и 332 адресов для подачи на буфер антенных выборок 172 адресов считывания.

Выходной сигнал накопителя сдвига 310 также сравнивается компаратором 326 с выходным сигналом счетчика системного времени 342 для формирования строба сдвинутого символа Уолша, который указывает, что буфер 172 антенных выборок имеет достаточно правильных данных для начала процесса поиска.

Счетчик поисковой группы 320 отслеживает количество элементов, которое осталось обработать в текущей поисковой группе. Счетчик поисковой группы 320 загружается по ширине поискового окна, записанной в буфер 314 ширины поиска в начале поискового окна. Счетчик поисковой группы 320 увеличивает свое значение, после того как завершится процесс накопления последнего символа Уолша каждой поисковой группы. Когда будет достигнуто конечное значение счетчика, это значит, что все сдвиги в поисковом окне обработаны. Для обеспечения индикации о том, что близок конец текущего поискового окна выходной сигнал счетчика поисковой группы 320 суммируется с помощью сумматора 324 с выходным сигналом буфера 312 ширины поисковой группы. Индикация окончания поискового окна показывает время, с которого буфер 172 антенных выборок может начать заполняться выборками данных от другой антенны при подготовке к следующему поисковому окну без потери содержимого, необходимого для текущего поискового окна.

Когда микропроцессор канального элемента 136 задает параметры поискового окна, он может установить, что поисковое окно будет выполняться для множества антенн. В таком случае идентичные параметры поискового окна повторяются с использованием выборок от ряда антенн. Такая группа поисковых окон называется набором антенных поисков. Если набор антенных поисков определяется микропроцессором канального элемента 136, то этот антенный набор программируется с помощью значения, записанного в буфере выбора антенны 348. После завершения набора антенных поисков микропроцессор канального элемента 136 приводится в состояние готовности.

Счетчик элементов 318 поисковой группы содержит количество элементов, оставшихся для обработки в текущей поисковой группе.

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Значение счетчика элементов возрастает на единицу с каждым обработанным элементом, и этот счетчик загружается выходным сигналом буфера 312 ширины поисковой группы, когда процесс поиска находится в состоянии ожидания или при завершении поисковой группы.

Счетчик накопления символов Уолша 322 подсчитывает количество символов Уолша, оставшихся для накопления в течение текущей поисковой группы. Этот счетчик загружается количеством символов Уолша для накопления, которое хранится в буфере 316 накопления символов Уолша, когда процесс поиска находится в состоянии ожидания или после завершения развертки поисковой группы при накоплении последнего символа Уолша. В противном случае счетчик при завершении каждой поисковой группы уменьшает свое значение.

Счетчик достоверности ввода 302 загружается всякий раз, когда изменяется входная антенна или настройка фильтра прореживания. Он загружается минимальным числом выборок, требуемым в процессе поиска для обработки поисковой группы, на основе выходного сигнала буфера 312 ширины поисковой группы /то есть одного символа Уолша плюс эквивалент ширины одной поисковой группы выборок/. Каждый раз, когда в буфер 172 антенных выборок записывается антенная выборка, значение счетчика достоверности ввода 302 возрастает на единицу. Когда счетчик достигает конечного значения, он посылает сигнал включения, разрешающий начало процесса поиска. Счетчик достоверности ввода 302 также обеспечивает механизм поддержания процесса поиска, когда сдвиги последующих поисковых окон не позволяют обеспечить непрерывную обработку данных.

Процесс поиска может находиться либо в состоянии ожидания, либо синхронизации, либо в активном состоянии. Блок управления упорядочением поиска 350 поддерживает текущее состояние. Интегральный поисковый процессор 128 использует состояние ожидания, когда на модем канального элемента 110 подается сигнал сброса. Во время состояния ожидания все счетчики и накопители в блоке управления поиском 166 загружаются соответствующими параметрами поиска, как было описано выше. Как только микропроцессор канального элемента 136 с помощью управляющего слова дает команды процессу поиска начинать непрерывный или одношаговый поиск, интегральный поисковый процессор 128 переходит в состояние синхронизации.

В состоянии синхронизации процесс поиска всегда ожидает границу сдвинутого символа Уолша. Если данные в буфере антенных выборок еще недостоверны или если установлен бит настройки группы управления мощностью и символ Уолша находится не на границе группы управления мощностью, то тогда интегральный поисковый процессор 128 остается в состоянии синхронизации, пока не возникнут соответствующие условия на границе следующего сдвинутого символа Уолша. При наличии соответствующего сдвинутого символа Уолша процесс поиска может перейти в активное состояние.

Интегральный поисковый процессор 128 остается в активном состоянии, пока он не обработает поисковую группу, и в этот момент он возвращается в состояние синхронизации. Если интегральный поисковый процессор 128 находится в одношаговом режиме, он может перейти из активного состояния в состояние ожидания после завершения последнего элемента поисковой группы для окончательного накопления символов Уолша для последней поисковой группы в поисковом окне. Затем интегральный поисковый процессор 128 ожидает команды от микропроцессора канального элемента 136, чтобы инициировать другой поиск. Если же интегральный поисковый процессор 128 находится в непрерывном режиме, то тогда в этот момент он загружает новый набор параметров поиска и возвращается в состояние синхронизации для ожидания сдвинутого символа Уолша с начальным сдвигом, подлежащего обработке при новом поиске. Активное состояние - это единственное состояние, в котором обрабатываются выборки антенных данных. В состояниях ожидания или синхронизации процесс поиска просто отслеживает время с помощью счетчика системного времени 342 и продолжает запись в буфер 176 ПСП и буфер 172 антенных выборок, так что, когда процесс поиска переходит в активное состояние, эти буферы будут готовы для использования.

На фиг. 14 в качестве примера показана временная диаграмма накопления первого символа Уолша второй поисковой группы в поисковом окне, например, в виде поисковой группы, показанной на фиг. 9. Третий символ Уолша, называемый опорным системным тактовым импульсом с нулевым сдвигом, показан разделенным на тридцать два интервала времени. Состояние поиска 372 изменяется от состояния синхронизации до активного, когда индикация границы сдвинутого символа Уолша, соответствующая символу Уолша 3, указывает, что буфер 172 антенных выборок готов с достоверными выборками к обработке на данном сдвиге. Во время следующего интервала времени обрабатывается первый элемент поисковой группы. Процесс поиска продолжается с использованием каждого интервала времени для обработки элемента поисковой группы, как показано символом "S" в интервалах времени 374, если входной каскад демодулятора 122 не использует процессор БПА 120, что показано символом "D" в интервалах времени 374. Процесс поиска заканчивает обработку каждого элемента в поисковой группе и возвращается в состояние синхронизации перед границей следующего сдвинутого символа Уолша, соответствующего символу Уолша 4. Также показано состояние счетчика поисковой группы 362, возрастающее в активном состоянии, пока оно не достигнет конечного состояния, указывающего на то, что обработана вся поисковая группа. Здесь показано возрастающее состояние счетчика сдвига 364 между интервалами времени, соответствующими элементу поисковой группы, так что это можно использовать для получения адреса считывания сдвига буфера выборки в течение интервала времени. Состояние счетчика сдвига 364 конвейерно задерживается в виде подсчета сдвига для регистра лучших результатов 366. Счетчик

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сдвига 368 получает приращения в процессе накопления конечного символа Уолша 370.

Таким образом, конфигурация однокристального поискового процессора благодаря буферизации антенных выборок и использованию процессора преобразования с квантованием времени может независимо устанавливать последовательность поиска, определяемую набором параметров поиска, анализировать результаты и представлять суммарный отчет о лучших траекториях, чтобы использовать их для повторного расширения элемента демодуляции. Это уменьшает относительную нагрузку на управляющий микропроцессор, так что можно использовать более дешевый микропроцессор, а также уменьшает непосредственные затраты на интегральные микросхемы, давая возможность выполнить весь модем канального элемента на одной микросхеме.

Описанные здесь общие принципы могут быть использованы в системах, где применяются альтернативные схемы передачи. Вышеуказанное описание основывалось на приеме сигнала обратного тракта, где отсутствует пилот-сигнал. По прямому тракту согласно предпочтительному варианту базовая станция передает пилот-сигнал. Пилот-сигнал - это сигнал, несущий известные данные, поэтому отпадает необходимость в процедуре БПА, используемой для определения того, какие данные были переданы. Для воплощения настоящего изобретения интегральный поисковый процессор для приема сигнала, содержащего пилот-сигнал, не содержит процессор БПА и не использует функцию обнаружения максимума. Процессор БПА и детектор максимума 160 на фиг. 5 могут быть заменены, например, простым накопителем 125, показанным на фиг. 15. Операция поиска при наличии пилот-сигнала аналогична операции поиска канала доступа в режиме захвата, описанном выше.

Вышеописанная архитектура поиска может быть использована для выполнения поисков самыми разными способами. Наиболее эффективным является линейный поиск. Линейный поиск выполняется путем линейного поиска потенциальных временных сдвигов в порядке, не зависящем от вероятности того, что удаленное устройство радиосвязи ведет передачу. При поиске сигнала удаленного устройства радиосвязи базовая станция должна знать ожидаемый диапазон зоны действия. Например, в предпочтительном варианте обычная базовая станция перекрывает диапазон порядка 50 километров, что подразумевает наличие задержки, связанной с подтверждением приема, 350 микросекунд или примерно 430 элементов псевдошумового кода. Также в среде многолучевого распространения, где сигналы имеют не прямые траектории, сигнал удаленного устройства радиосвязи может быть задержан чуть ли не вдвое по сравнению с прямой траекторией распространения, имея в виду, что поиск должен вестись по набору из почти 1000 различных сдвигов псевдошумового сигнала. При обнаружении сигнала удаленного устройства радиосвязи он демодулируется и становится известным примерное расстояние до удаленного устройства радиосвязи. В связи с этим возможные сдвиги псевдошумового сигнала, которые необходимо определить, чтобы обеспечить обнаружение большинства достоверных многолучевых сигналов, существенно уменьшаются.

В рамках данного поиска по группе управления мощностью имеются три причины, по которым сигнал не может быть обнаружен при данном сдвиге псевдошумового сигнала. Во-первых, сигнал может не прийти с заданным сдвигом. Удаленное устройство радиосвязи может выдать несколько сигналов многолучевого распространения, но количество создаваемых многолучевых сигналов составляет лишь очень малую часть от всех сдвигов, которые подвергаются поиску. Таким образом, большинство сдвигов превышают порог обнаружения, потому что нет сигнала удаленного устройства радиосвязи с данным сдвигом.

Во-вторых, сигнал может поступать с заданным сдвигом псевдошумового сигнала, но с замиранием на протяжении большей части всего времени поиска. Как пояснялось выше, характеристики многолучевого распространения радиоканала могут привести к замиранию сигнала. Замирание определяется характеристиками фазирования канала с многолучевым распространением. Замирание появляется, когда векторы многолучевого распространения суммируются неблагоприятным образом, образуя принимаемый сигнал, по уровню меньший, чем любой отдельный вектор. Таким образом, если сигнал, который долгое время был достоверным, вдруг сильно замирает во время проведения поиска, то он не сможет быть обнаружен в процессе поиска.

В третьих, сигнал может поступить с заданным сдвигом псевдошумового сигнала, но в случае, когда передатчик удаленного устройства радиосвязи в рассматриваемый период времени не передает сигнал. Как пояснялось выше, в предпочтительном варианте удаленное устройство радиосвязи формирует пакетный сигнал. Удаленное устройство радиосвязи содержит вокодер с регулируемой частотой, который формирует блоки данных с изменяемой частотой. Рандомизатор пакета данных определяет, в течение каких периодов времени удаленное устройство радиосвязи ведет передачу и в течение каких периодов времени он не ведет передачу данных, выдавая скорость передачи данных сигнала, подлежащего передаче, конкретный идентификационный номер удаленного устройства радиосвязи и время суток. При работе с частотой меньшей полной частоты, рандомизатор пакета данных в удаленном устройстве радиосвязи распределяет случайным образом периоды активного времени внутри пакета передачи. Соответствующий рандомизатор пакета данных также включается в состав базовой станции, так что базовая станция может воссоздать псевдослучайное распределение на основе времени суток и конкретного идентификационного номера удаленного устройства радиосвязи, но во время процесса поиска отсутствует информация о скорости передачи. Как было отмечено выше, периоды времени восьмикратной частоты определяют так называемую совершенную группу временных интервалов. Таким путем, независимо от скорости данных в

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передаваемом сигнале, каждый временной период, соответствующий совершенной группе, точно соответствует интервалу времени, когда соответствующее удаленное устройство радиосвязи передавало сигнал. Во время всех других временных интервалов удаленное устройство радиосвязи может передавать или не передавать данные в зависимости от соответствующей скорости кодирования.

Если установлен линейный поиск, для того чтобы получить достоверные измерения мощности, то процесс поиска ограничивает полное время поиска /то есть количество накоплений символов Уолша на одном поисковом сдвиге/, чтобы начать и закончить поиск в границах одной группы управления мощностью, как более подробно объяснялось выше. Считается, что поиск, идущий только внутри одной группы управления мощностью, синхронизирован с границами группы управления мощностью. Если процесс поиска с данным сдвигом накапливался безотносительно границ группы управления мощностью и удаленное устройство радиосвязи передавало со скоростью, меньшей полной скорости, то достоверные результаты поиска, соответствующие группе управления мощностью, где стробировался сигнал удаленного устройства радиосвязи, могут быть просуммированы с шумом, накопленным во время следующей группы управления мощностью, когда сигнал от удаленного устройства радиосвязи не проходил. Суммирование результатов поиска, соответствующих группе управления мощностью, когда сигнал удаленного устройства радиосвязи не проходил, разрушает полезные результаты, накопленные во время управления мощностью, когда сигнал удаленного модуля стробировался.

Один из способов поиска можно использовать для поиска только тех групп управления мощностью, которые соответствуют совершенным группам. Даже если выполняется поиск только совершенной группы, процесс поиска и процесс распределения элементов демодуляции должен быть в состоянии обработать ситуацию, при которой накопленная энергия не превышает порог обнаружения, но в действительности сигнал со сдвигом присутствует благодаря характеристикам непредсказуемого замирания канала. Таким образом, есть более эффективная схема для накопления энергии во всех группах управления мощностью, независимо от того, соответствуют или нет они совершенным группам. Если при поиске обнаруживается энергия, которая не соответствует совершенной группе, генерируются дополнительные достоверные данные в добавление к данным, генерируемым на основе поиска только совершенной группы.

Как отмечалось выше, поиск преамбулы отличается от поиска, выполняемого во время работы канала трафика. Когда удаленное устройство радиосвязи изначально пытается получить доступ в систему, он посылает сигнал радиомаяка, называемый преамбулой, в котором используется нулевой символ Уолша. Нулевой символ Уолша - это символ Уолша, который содержит все логические нули вместо половины единиц и половины нулей, как описывалось выше. При выполнении поиска преамбулы поисковое устройство ищет любое удаленное устройство радиосвязи, посылающее сигнал радиомаяка из нулевых символов Уолша по каналу доступа. В предпочтительном варианте передача преамбулы всегда идет с полной скоростью и никогда не прерывается. Таким образом, во время поиска преамбулы нет необходимости в синхронизации с границами группы управления мощностью.

Существует множество конфигураций систем связи с коллективным доступом и расширенным спектром, подробно здесь не описанных, но в которых применимо настоящее изобретение. Например, вместо кодирования Уолша и декодирования с использованием БПА можно было бы использовать другие методы кодирования и декодирования. Предыдущее описание предпочтительных вариантов предназначено для того, чтобы специалисты могли осуществить и использовать настоящее изобретение. Специалистам должны быть очевидны различные модификации этих вариантов, а сформулированные здесь исходные принципы могут быть применены в других вариантах без использования изобретательских способностей. Таким образом, предполагается, что настоящее изобретение не ограничивается показанными здесь вариантами его осуществления, а должно соответствовать самому широкому объему, согласующемуся с раскрытыми здесь принципами и новыми признаками.

Формула изобретения:

1. Способ приема сигнала, содержащего группу сигналов вызовов с расширенным спектром, совместно использующих общую полосу частот, в котором каждый из указанных сигналов вызовов с расширенным спектром содержит последовательность бит, закодированных в группах фиксированной длины в виде последовательности символов, причем последовательность указанных символов группируется вместе в группы управления, мощностью, где каждый символ в общей группе управления мощностью передается с общим уровнем мощности, и указанные группы управления мощностью передаются в пакетах, и выделения одного из указанных сигналов вызова из указанной группы для определения уровня сигнала вызова с временным сдвигом из-за задержки на трасре распространения по отношению к опорному времени с нулевым сдвигом, отличающийся тем, что способ включает этапы, при которых осуществляют запоминание бит данных псевдошумовой последовательности (ПСП) в буфере ПСП, запоминание первого принятого набора выборок сигналов вызовов в буфере выборок, имеющем ограниченную емкость, сжатие набора первой фиксированной длины из указанных выборок сигналов вызовов из буфера выборок, соответствующих первому времени задержки на трассе распространения, с помощью первого набора бит данных ПСП из буфера ПСП для получения первого сжатого выходного сигнала, запоминание второго принятого набора выборок сигналов вызовов в буфере выборок, сжатие набора второй фиксированной длины из выборок сигналов вызовов из буфера выборок,

соответствующих второму времени задержки на трассе распространения, с помощью первого набора бит данных ПСП из указанного буфера ПСП для получения второго сжатого выходного сигнала, при этом набор второй фиксированной длины из выборок сигналов вызовов содержит большое число таких же выборок сигналов вызовов, что и набор первой фиксированной длины из выборок сигналов вызовов, длина первого и второго принятых наборов выборок сигналов вызовов является частью фиксированной длины набора первой и второй фиксированной длины из выборок сигналов вызовов, причем этапы запоминания набора первой и второй фиксированной длины из выборок сигналов вызовов и этапы сжатия набора первой и второй фиксированной длины из выборок сигналов вызовов выполняются независимо от вероятности того, что один из сигналов вызовов содержит одну из групп управления мощностью.

2. Способ приема сигнала, содержащего группу сигналов вызовов с расширенным спектром, совместно использующих общую полосу частот, и выделения первого сигнала из группы сигналов с расширенным спектром для определения уровня сигнала с временным сдвигом из-за задержки на трассе распространения по отношению к опорному времени с нулевым сдвигом первого сигнала, причем первый сигнал содержит последовательность символов, которая группируется вместе в набор символов, каждый символ в общем наборе символов передается с фиксированным уровнем мощности, а последовательные наборы символов могут передаваться с различными уровнями сигнала, причем указанные различные уровни сигнала включают в себя нулевой уровень, когда передача первого сигнала прерывается, отличающийся тем, что включает этапы, при которых осуществляют поиск первого набора выборок сигналов вызовов, соответствующего первому набору символов для первого сигнала с первым сдвигом для получения первой оценки его мощности, поиск второго набора выборок сигналов вызовов, соответствующих первому набору символов для первого сигнала с первым сдвигом для получения второй оценки его мощности, суммирование первой и второй оценок мощности для получения оценки уровня мощности набора символов с первым сдвигом, поиск третьего набора выборок сигналов вызовов, соответствующего второму набору символов для первого сигнала со вторым сдвигом для получения третьей оценки его мощности, поиск четвертого набора выборок сигналов вызовов, соответствующего второму набору символов для первого сигнала с вторым сдвигом для получения четвертой оценки его мощности и суммирование третьей и четвертой оценок мощности для получения оценки уровня мощности набора символов с вторым сдвигом, при этом первый набор символов и второй набор символов соответствуют наборам смежных во времени символов, причем этапы поиска выполняются непрерывно независимо от фиксированного уровня мощности.

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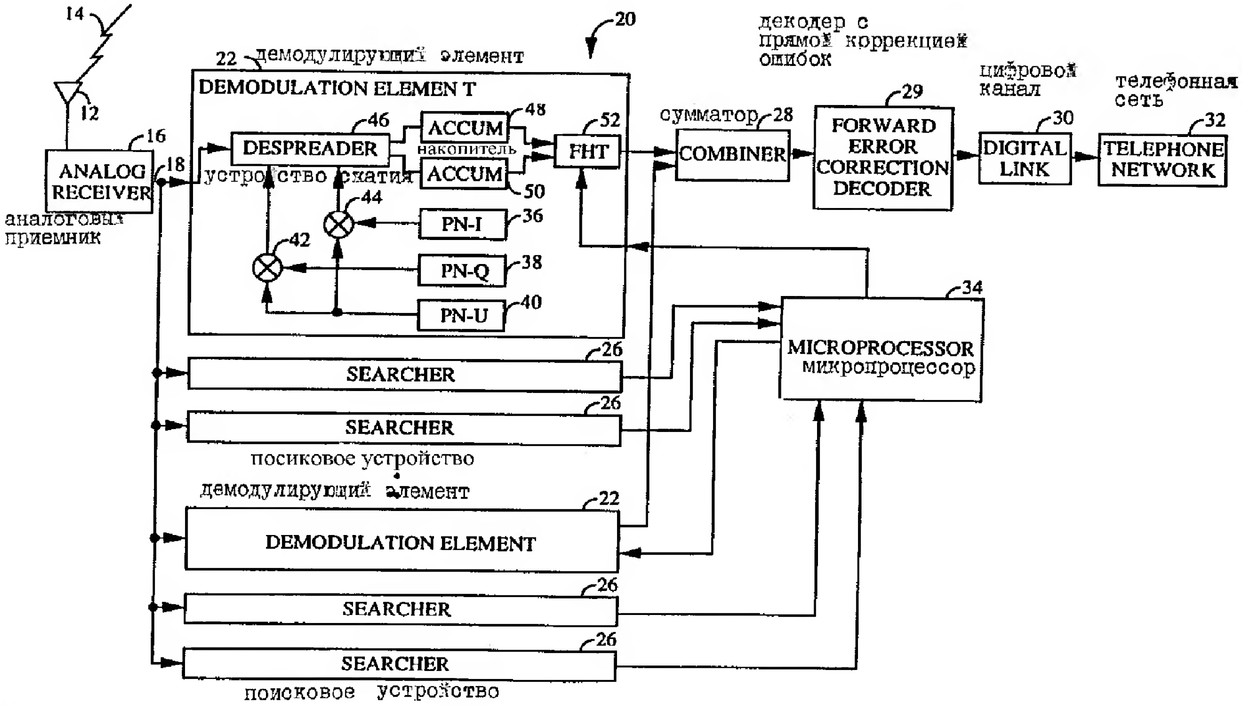
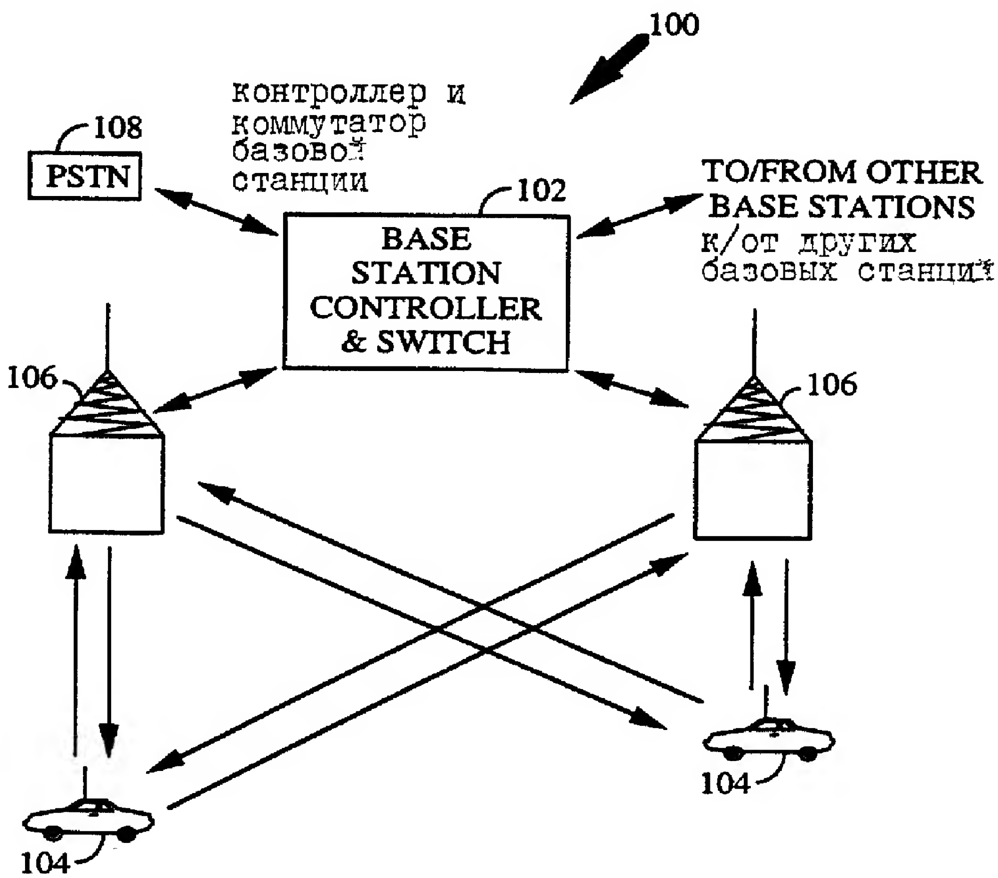


Фиг.2



Фиг.3

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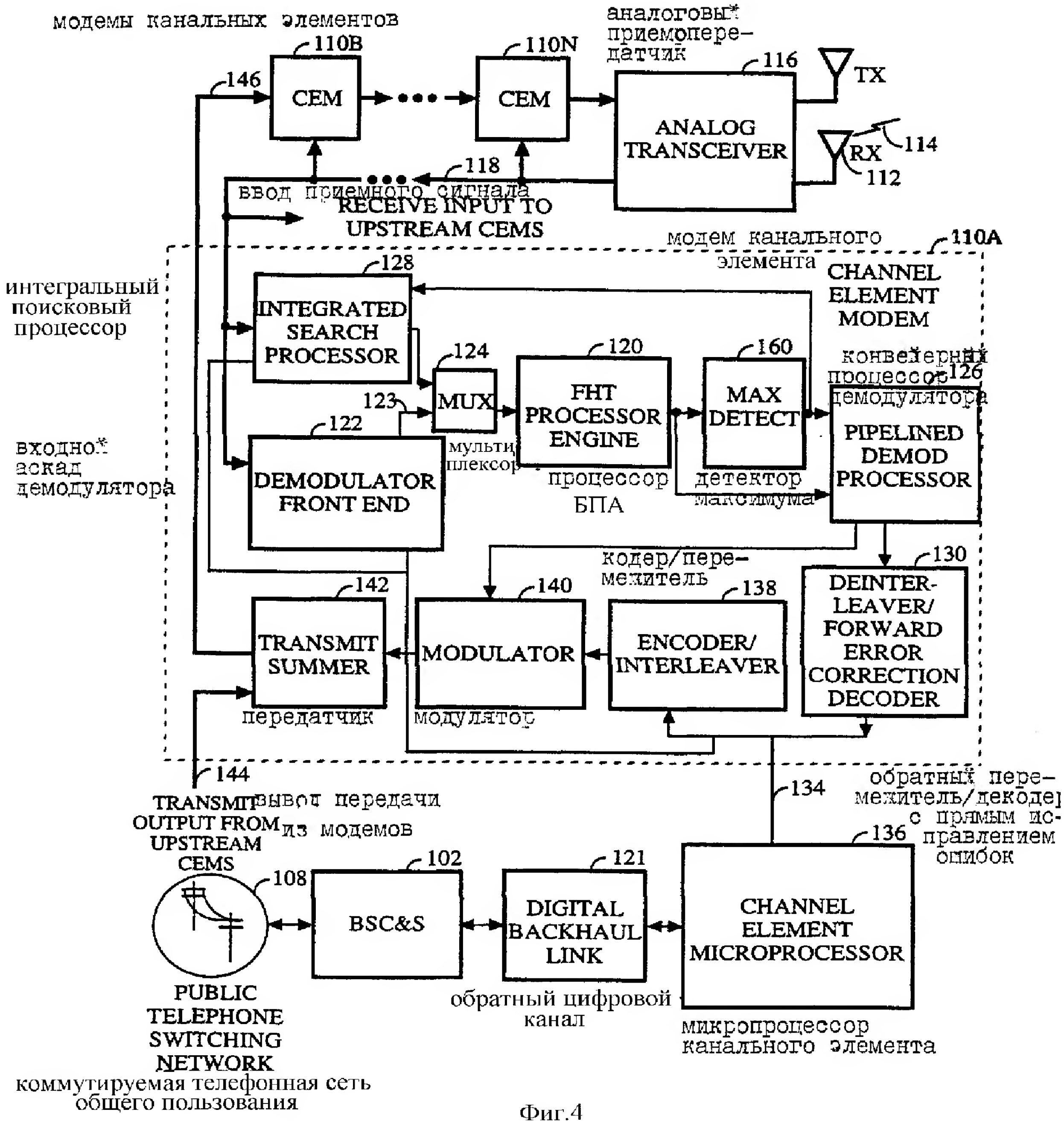
RU 2157592 C2



Фиг.4

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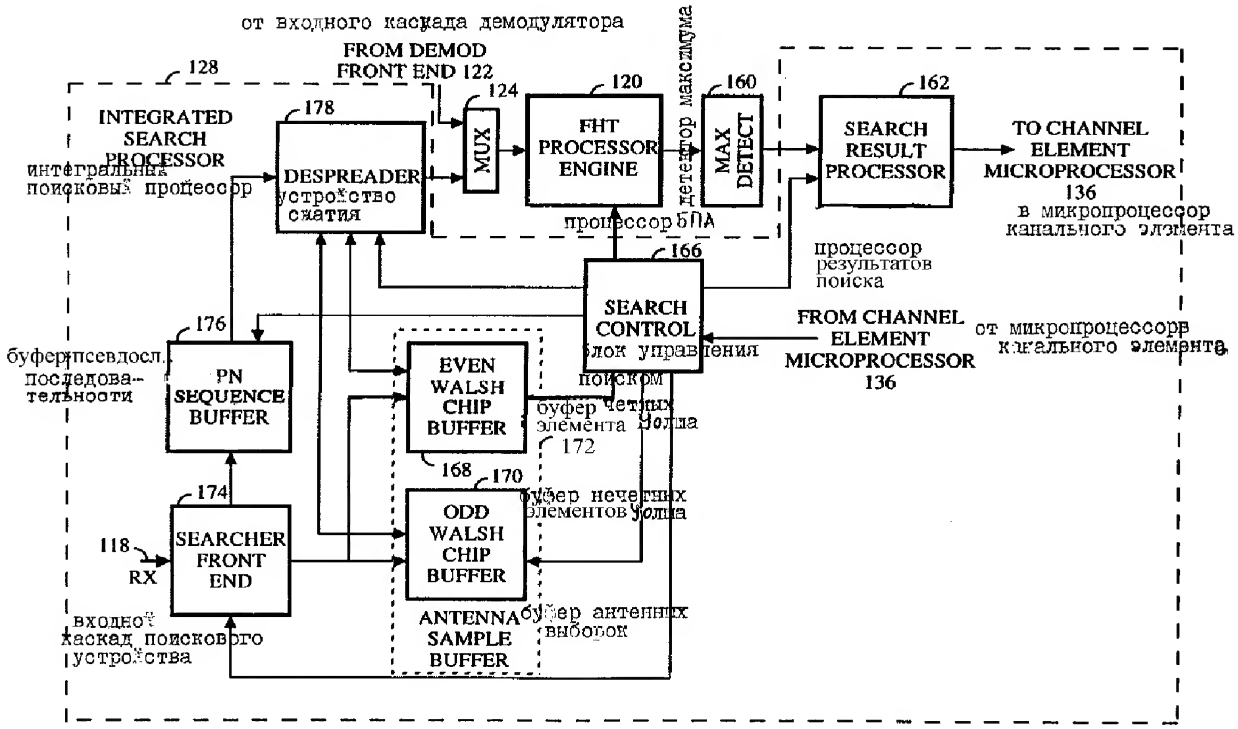

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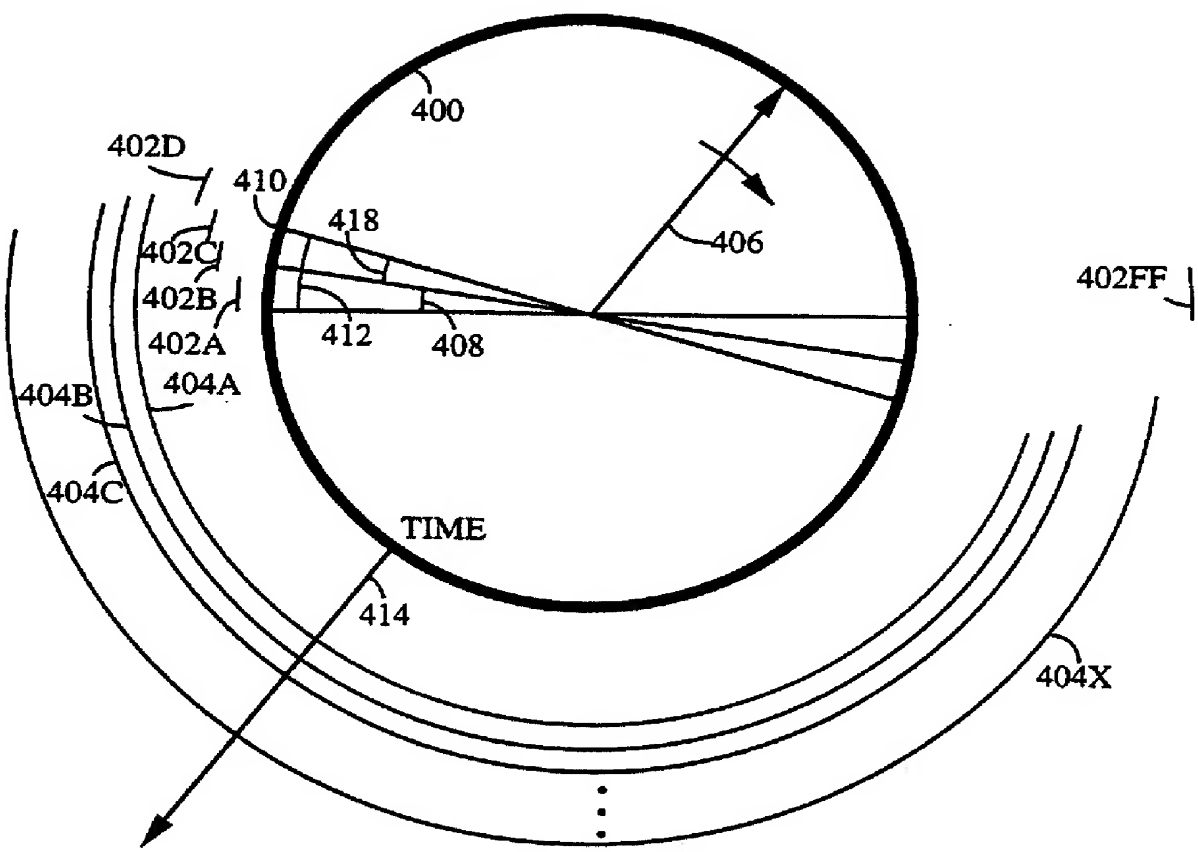

Фиг.6

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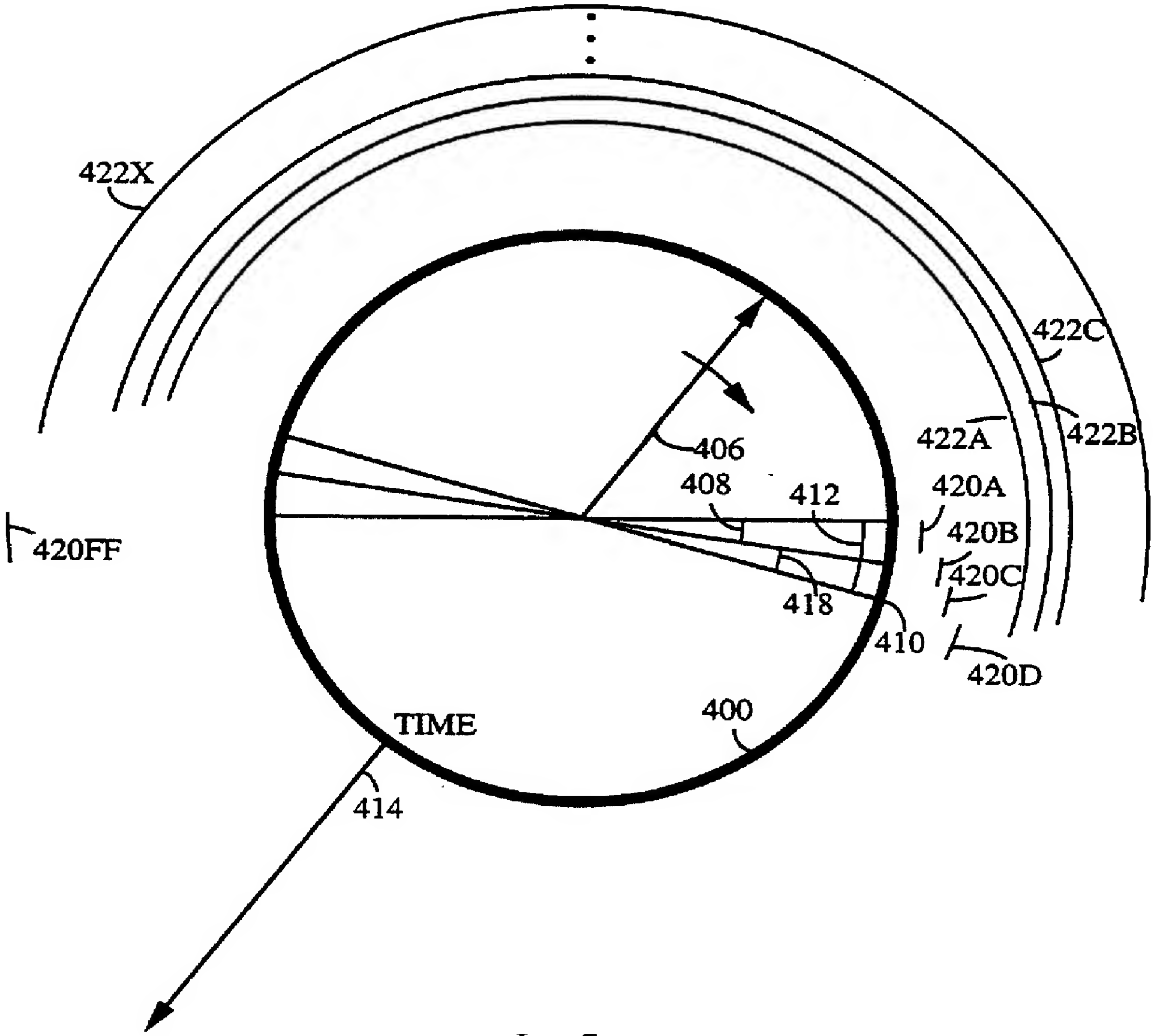
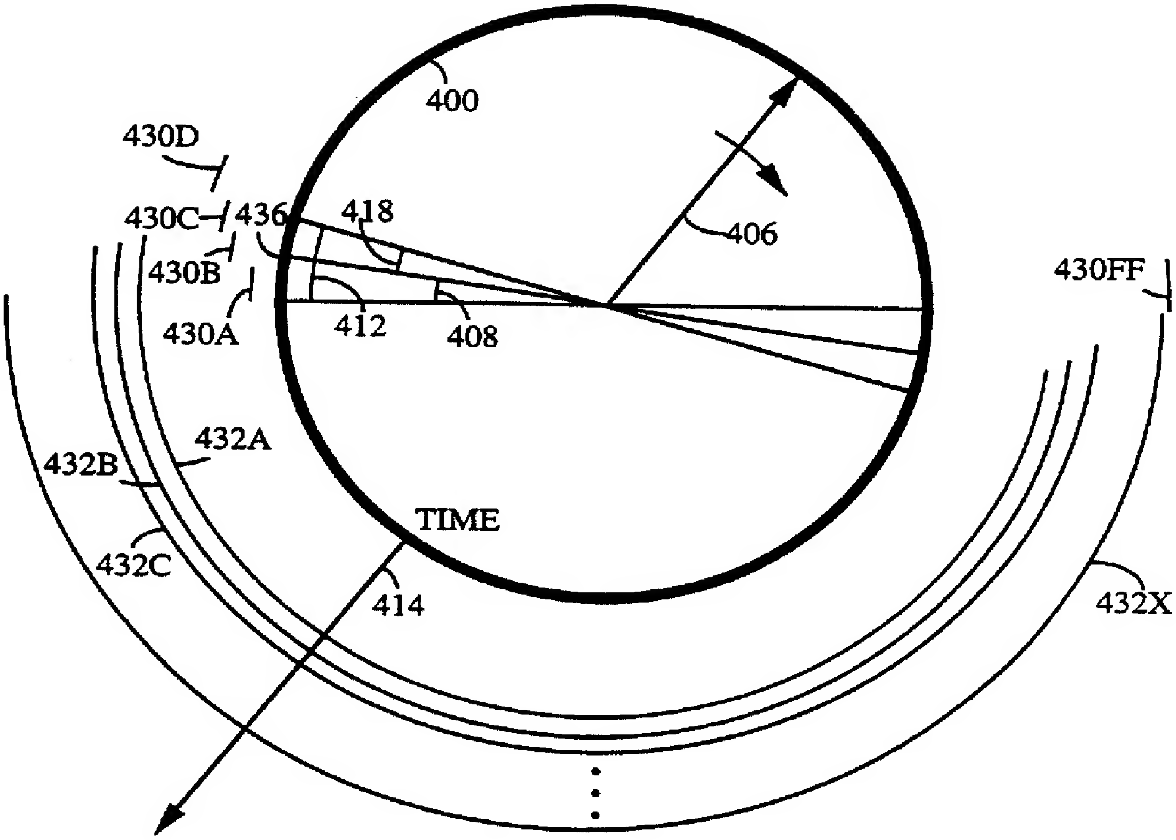


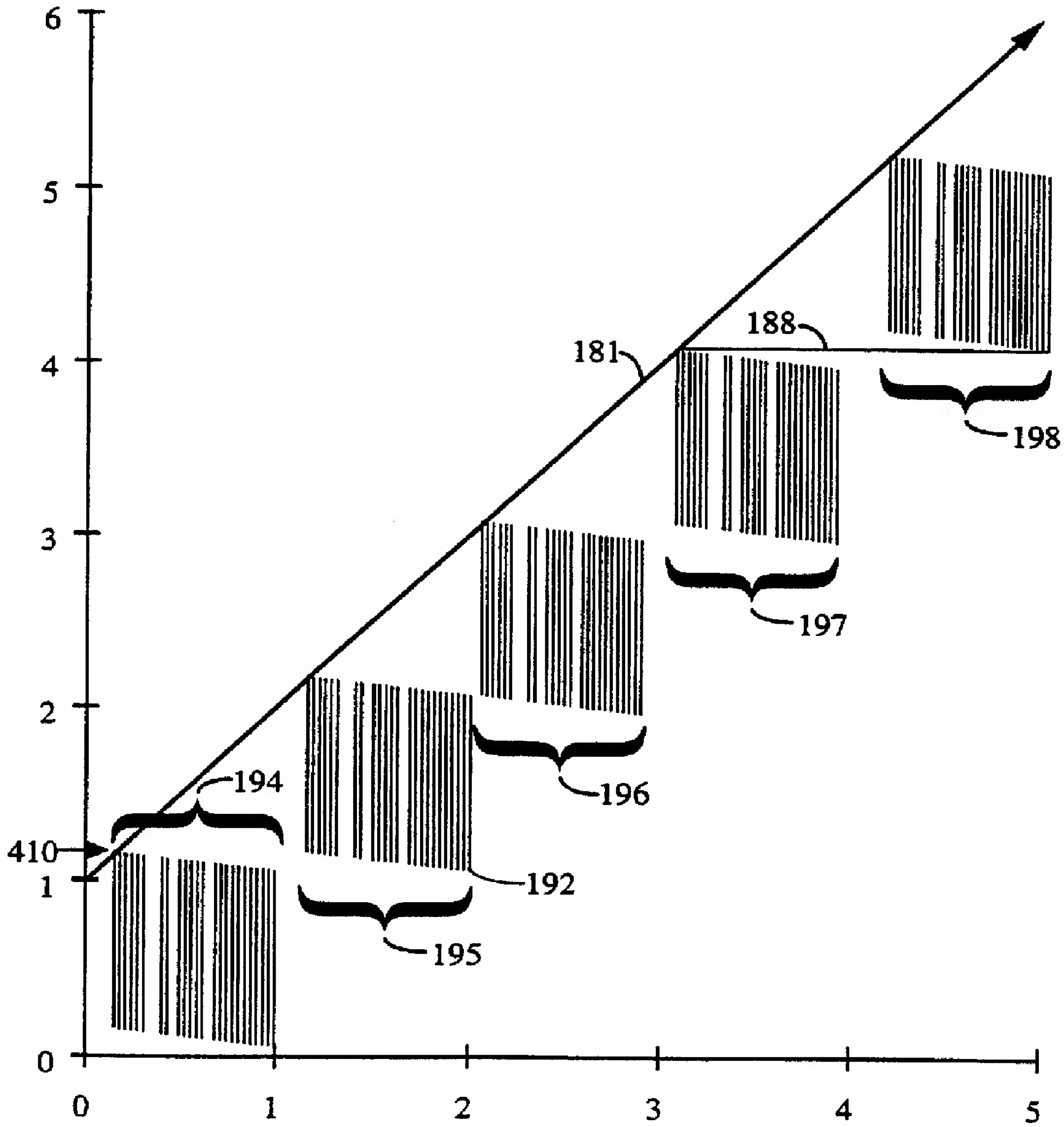
Фиг.7

RU 2157592 C2



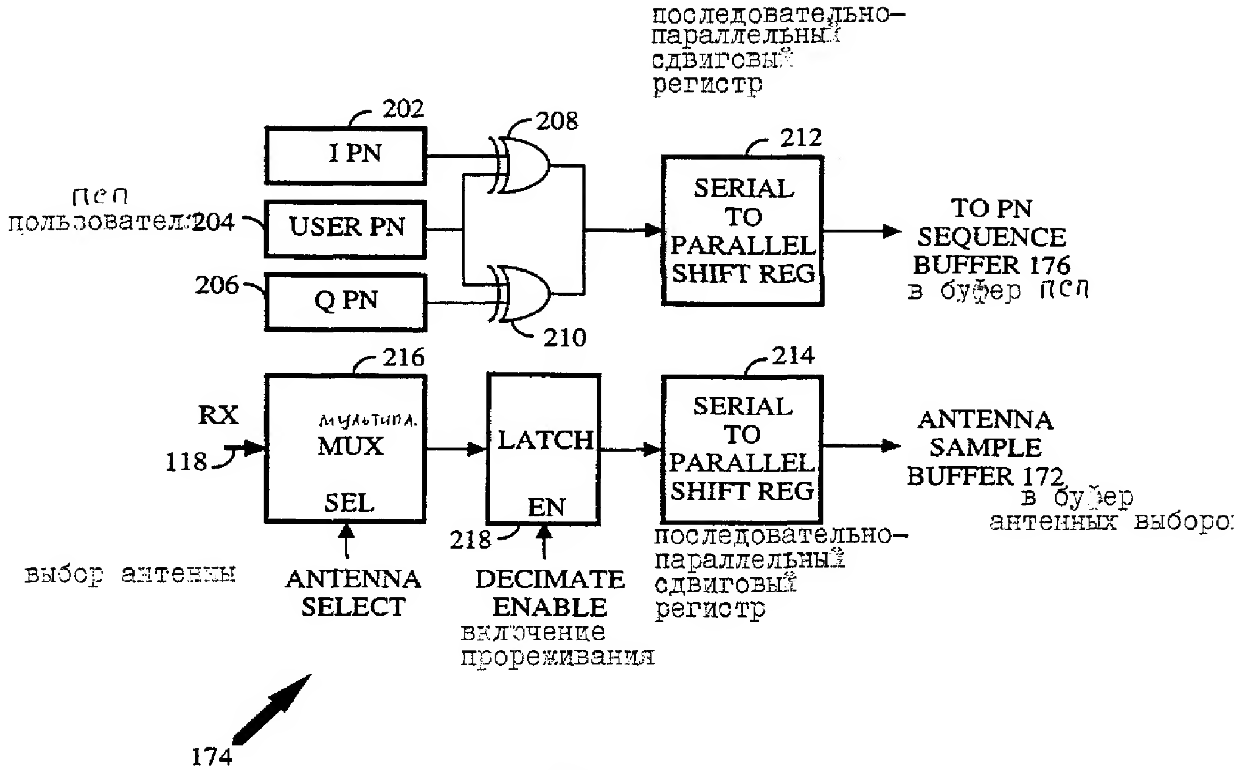
Фиг.8

RU 2157592 C2

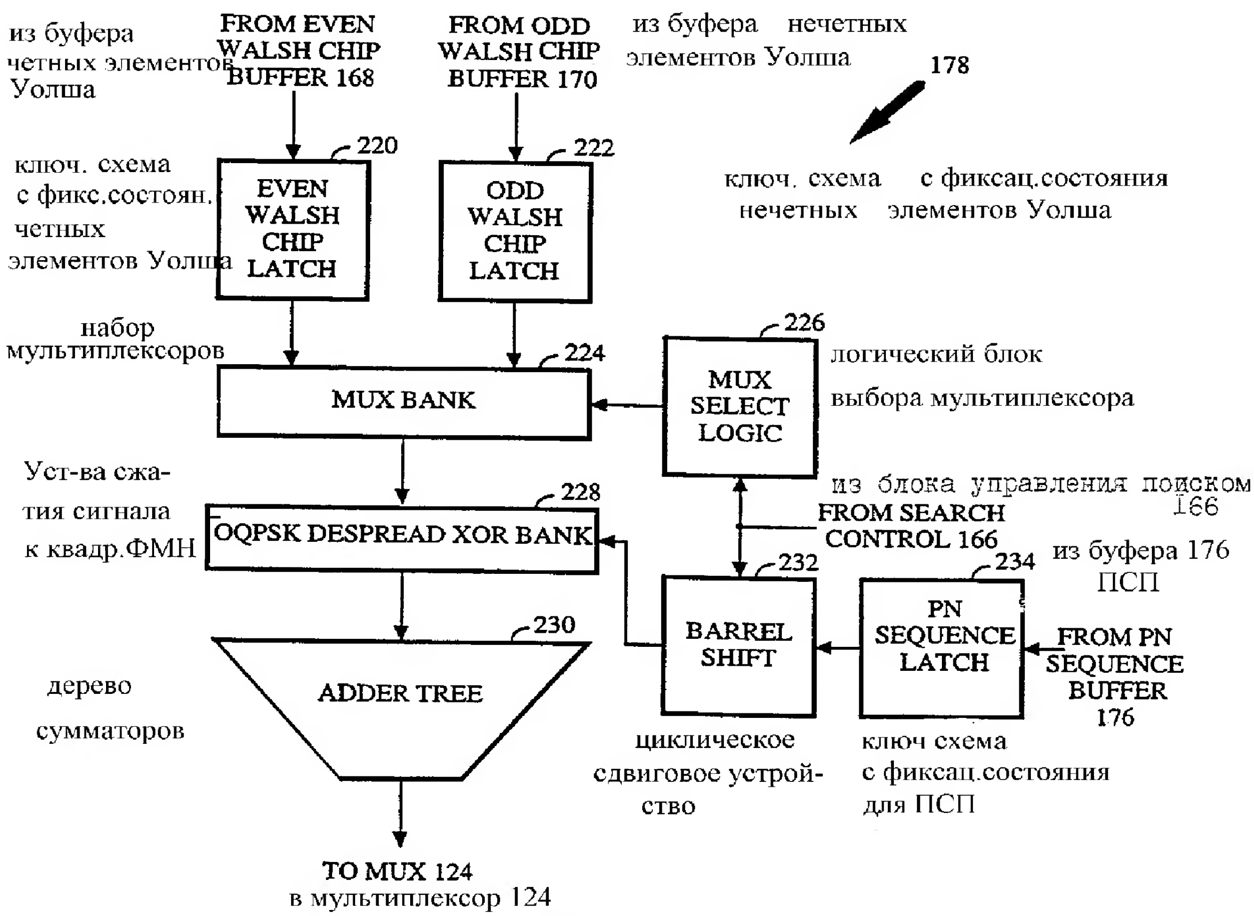


Фиг.9

RU 2157592 C2



Фиг.10



Фиг.11

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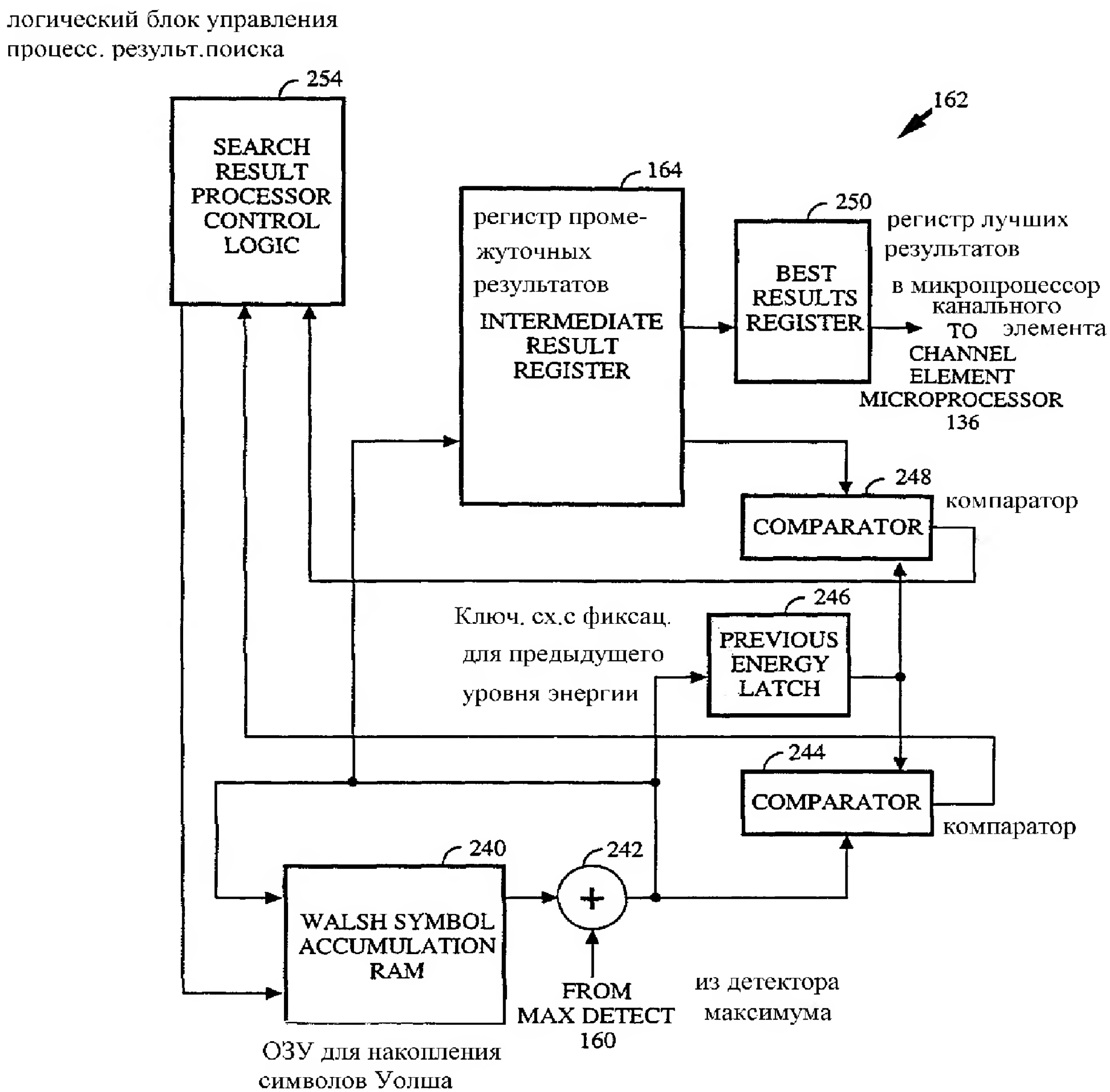

Фиг.12

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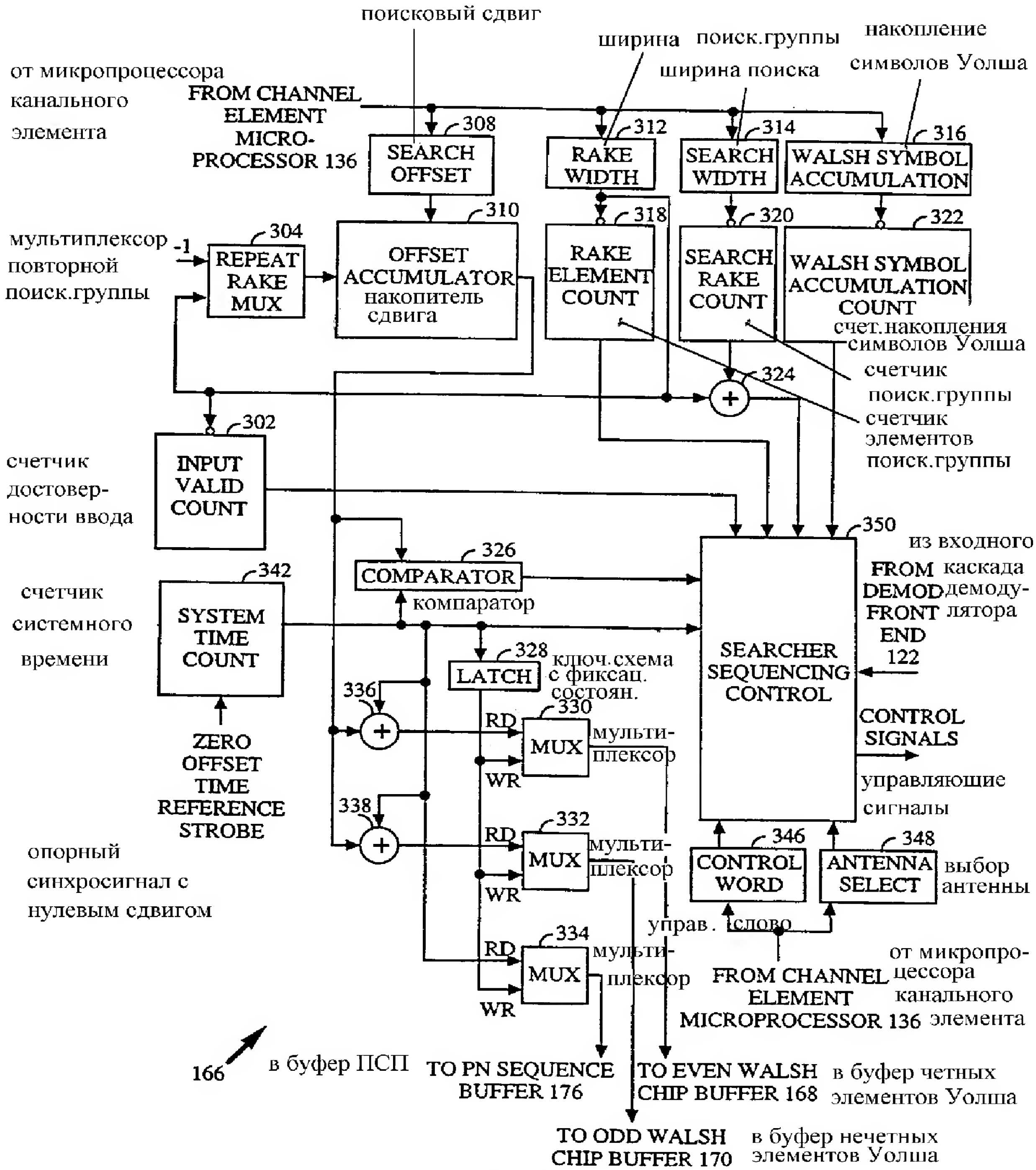

Фиг.13

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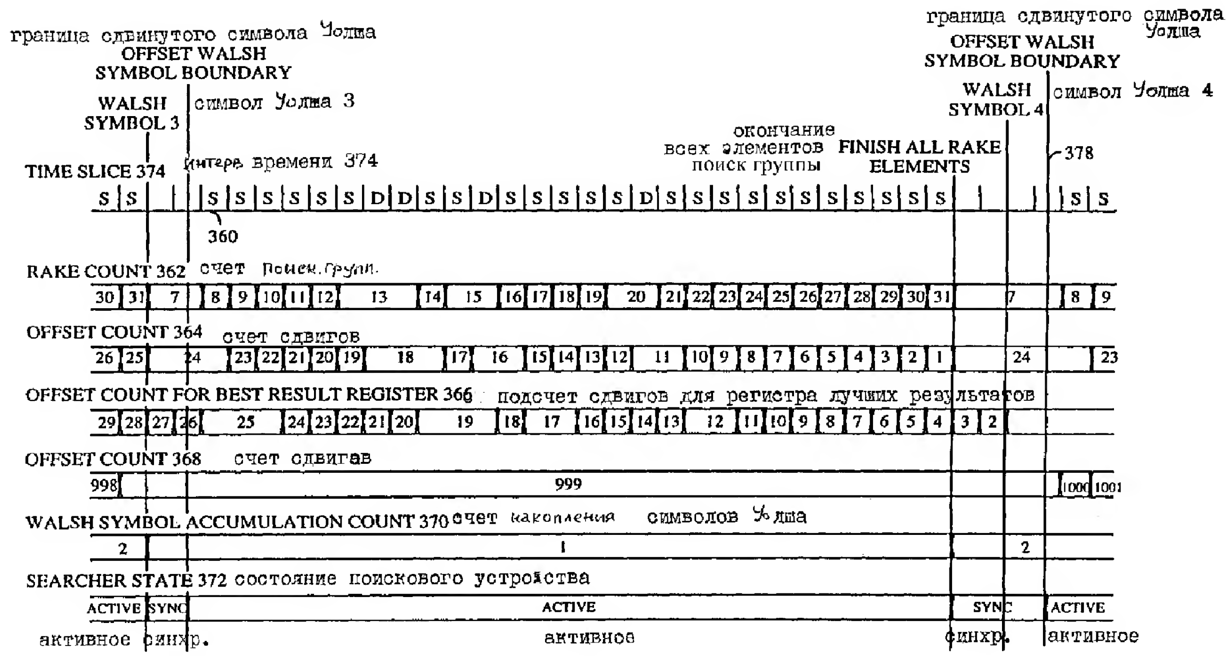
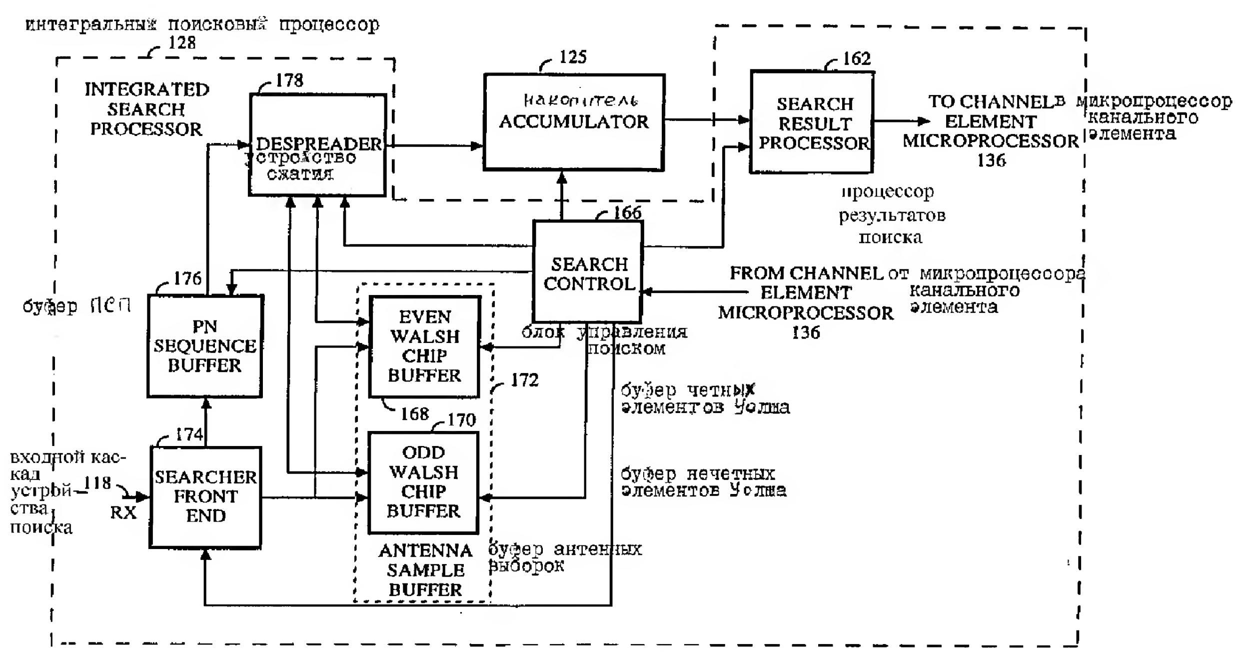


Фиг.14



Фиг.15

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Patent/Publication Number	419912
Title	Fast fourier transform based CDMA RAKE receiver structure and method
Issued/Publication Date	2001/01/21
Application Date	1999/01/25
Application Number	088101059
Certification Number	127553
IPC	H04B-001/707
Inventor	HUANG, JIA-QI TW; WANG, XIN-YUAN TW; HUANG, YONG-LIANG TW
Applicant	INDUSTRIAL TECHNOLOGY RESEARCH INSTITUTETW
Priority Number	19980811 NA NA
Abstract	The present invention provides a CDMA RAKE receiver which uses Fast Fourier Transform (FFT) matched filter to calculate a data detection. The received signal is processed by the RAKE receiver in the frequency domain. The RAKE receiver comprises a pilot signal spreading code matched filter, a data signal spreading code matched filter and a channel matched filter. The pilot signal spreading code matched filter removes a spreading code of the pilot signal. The data signal spreading code matched filter removes a multiple access spreading code of the data signal. A channel matched filter estimates the channel frequency response and combines the received data signal from different paths and generates a decision. In order to increase the capacity of CDMA system, the RAKE receiver employs an interference cancellation method. A downlink receiver on a mobile station can estimate the interference of a pilot signal and remove the pilot interference from the received data before the data detection. An uplink receiver on a base station employs a multi-stage parallel interference cancellation technique to remove the multiple access interference from other users.

中華民國專利公報 [19] [12]

[11]公告編號：419912

[44]中華民國　90 年 (2001)　01 月 21 日

發明　　全 10 頁

[51] Int.Cl ^{06}: H04B1/707

[54]名　稱：以快速傅利葉轉換為基礎之分碼多重進接犁耙式接收機架構及其方法

[21]申請案號：088101059　[22]申請日期：中華民國　88年 (1999)　01 月 25 日

[30]優 先 權：[31]09/132,859　[32]1998/08/11　[33]美國

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[74]代 理 人：

[57]申請專利範圍：

1.一種下鏈接收機，包括：

一領航干擾消除單元，用以從一接收基頻信號之快速傅利葉轉換中減去領航干擾信號；

一資訊信號展頻碼匹配濾波器，用以從該領航干擾消除單元之輸出中去除資訊信號展頻碼；

一領航信號展頻碼匹配濾波器，用以從該接收基頻信號之快速傅利葉轉換中去除領航信號展頻碼；

一通道匹配濾波器，以該領航信號展頻碼匹配濾波器之輸出為基礎產生一通道頻率響應估計，接著產生該資訊信號展頻碼匹配濾波器其輸出與該通道頻率響應估計之共軛複數之乘積，最後計算在一展頻碼週期中該乘積之積項和；以及

一判定單元，基於一預選調變形式以該通道匹配濾波器產生之積項和來判定該資訊值。

2.如申請專利範圍第1項所述之該下鏈接收機，其中該領航干擾消除單元包括：

一領航信號估計單元，用以從該接收基頻信號之快速傅利葉轉換中估計該領航信號成分；以及

一結合器，用以從該接收基頻信號之快速傅利葉轉換中減去該估計領航信號成分。

3.如申請專利範圍第2項所述之該下鏈接收機，其中該領航信號估計單元包括：

一延遲單元，延遲該接收基頻信號之快速傅利葉轉換一展頻碼週期；以及

一平均單元，用以執行該延遲後接收基頻信號之快速傅利葉轉換之一加權平均。

4.如申請專利範圍第2項所述之該下鏈接收機，其中該領航信號估計單元包括：

一延遲單元，延遲該接收基頻信號之快速傅利葉轉換一展頻碼週期；

一資訊信號重建單元，以資訊判定回授的概念為基礎重建該資訊信號成分；

一領航信號估計結合器，從該延遲後接收基頻信號之快速傅利葉轉換中減去該重建資訊信號；以及

一平均單元，用以執行該領航信號估計結合器其結果之一加權積項平均。

5.如申請專利範圍第1項所述之該下鏈接收機，其中該資訊信號展頻碼匹配濾波器包括：

一儲存單元，用以儲存一資訊信號展頻碼之快速傅利葉轉換；

一共軛複數單元，用以計算該資訊信號展頻碼其快速傅利葉轉換之共軛複數；

一乘法器，用以將該領航干擾消除單元之輸出與該共軛複數單元之輸出相乘。

6.如申請專利範圍第1項所述之該下鏈接收機，其中該領航信號展頻碼匹配濾波器包括：

一儲存單元，用以儲存一領航信號展頻碼之快速傅利葉轉換；

一共軛複數單元，用以計算該領航信號展頻碼其快速傅利葉轉換之共軛複數；

一乘法器，用以將該接收基頻信號之快速傅利葉轉換與該共軛複數單元輸出相乘。

7.如申請專利範圍第1項所述之該下鏈接收機，其中該通道匹配濾波器包括一通道頻率響應估計單元，以該領航信號展頻碼匹配濾波器之輸出為基礎產生該通道頻率響應估計，該通道頻率響應估計單元包括：

一延遲單元，延遲該領航信號展頻碼匹配濾波器之輸出一展頻碼週期；

一平均單元，用以執行該延遲後領航信號展頻碼匹配濾波器輸出之一加權平均；以及

一共軛複數單元，用以計算該延遲後領航信號展頻碼輸出加權積項平均的共軛複數。

8.如申請專利範圍第7項所述之該下鏈接收機，其中該通道頻率響應估計單元包括一保留主路徑單元用以保留複數主路徑，該保留主路徑單元包括：

一反快速傅利葉轉換單元(IFFT)，用以計算該延遲後領航信號展頻碼匹配濾波器輸出之加權平均之一反快速傅利葉轉換，產生一通道脈衝響應估計；

一搜尋單元，從該通道脈衝響應估計中搜尋出振幅峰值；

一選取單元，以該搜尋得之振幅峰值為基礎選取主要路徑，並產生一保留主要路徑之通道脈衝響應估計；以及

一快速傅利葉轉換單元，用以計算出該保留主要路徑之通道脈衝響應估計之快速傅利葉轉換。

9.如申請專利範圍第7項所述之該下鏈接收機，其中該通道匹配濾波器還包括：

一乘法器，將該資訊信號展頻碼匹配濾波器之輸出與該通道頻率響應估計相乘；以及

一加法器，計算在一展頻碼週期中該乘法器其輸出之一積項和。

10.一種下鏈接收的方法，包括：

從一接收基頻信號之快速傅利葉轉換中減去領航干擾信號；

該接收基頻信號之快速傅利葉轉換減去該領航干擾信號後，再去除一資訊信號展頻碼；

從該接收基頻信號之快速傅利葉轉換在去除一領航信號展頻碼；

該接收基頻信號之快速傅利葉轉換在去除該領航信號展頻碼後，用以估計

一通道頻率響應；

該接收基頻信號在減去領航干擾信號與去除該資訊信號展頻碼後，再與該通道頻率響應估計之共軛複數相乘後得一乘積；

計算該乘積在一展頻週期內之積項和；以及

基於一預選調變形式以該積項和值判定該資訊值。

11.如申請專利範圍第10項所述之該方法，其中從一接收基頻信號之快速傅利葉轉換中減去領航干擾信號，包括：

從該接收基頻信號之快速傅利葉轉換中估計該領航信號成分；以及

從該接收基頻信號之快速傅利葉轉換中減去該估計領航信號成分。

12.如申請專利範圍第11項所述之該方法，其中估計該領航信號成分包括：

延遲該接收基頻信號之快速傅利葉轉換一展頻碼週期；以及

執行該延遲後接收基頻信號之快速傅利葉轉換之一加權平均。

13.如申請專利範圍第11項所述之該方法，其中估計該領航信號成分包括：

重建該前一個接收符號之資訊信號成分；

延遲該接收基頻信號之快速傅利葉轉換一展頻碼週期；

從該延遲後接收基頻信號之快速傅利葉轉換中減去該重建資訊信號成分；以及

該延遲後接收基頻信號之快速傅利葉轉換在減去該重建資訊信號成分後計算其一加權平均。

14.如申請專利範圍第10項所述之該方法，其中該接收基頻信號之快速傅利葉轉換在減去該領航干擾信號後，再去除資訊信號展頻碼包括：

產生一資訊信號展頻碼之快速傅利葉轉換；

計算該資訊信號展頻碼其快速傅利葉轉換之共軛複數；以及

一接收基頻信號之快速傅利葉轉換在減去該領航干擾信號後，再與該計算之共軛複數相乘。

15.如申請專利範圍第10項所述之該方法，其中從該接收基頻信號之快速傅利葉轉換中去除領航信號展頻碼包括：

產生一領航信號展頻碼之快速傅利葉轉換；

計算該領航信號展頻碼其快速傅利葉轉換之共軛複數；以及

將該接收基頻信號之快速傅利葉轉換與該計算之共軛複數相乘。

16.如申請專利範圍第10項所述之該方法，其中估計通道頻率響應包括：

該接收基頻信號之快速傅利葉轉換在去除該領航信號展頻碼後，再延遲一展頻碼週期；

該接收基頻信號之快速傅利葉轉換在去除該領航信號展頻碼及延遲一展頻碼週期後，計算其一加權平均；以及

計算該加權平均之共軛複數。

17.如申請專利範圍第16項所述之該方法，其中估計通道頻率響應還包括保留複數主要路徑，保留複數主要路徑包括：

計算該加權平均之一反快速傅利葉轉換而產生一通道脈衝響應估計；

從該該通道脈衝響應中搜尋出該振幅峰值；

以該搜尋得之振幅峰值為基礎選取主要路徑而產生一保留主要路徑之通道脈衝響應估計；以及

計算該保留主要路徑之通道脈衝響應估計之快速傅利葉轉換。

18.如申請專利範圍第16項所述之該方法，其還包括：

該接收基頻信號之快速傅利葉轉換在減去領航信號與去除該資訊信號展頻碼後，與該通道頻率響應估計之共軛複數相乘；以及

計算該乘積在一展碼週期內之積項和。

19.一種下鏈接收機，包括：

一領航干擾消除單元，用以從一接收基頻信號之快速傅利葉轉換中減去領航干擾信號；

一資訊信號展頻碼匹配濾波器，用以從該領航干擾消除單元之輸出中去除資訊信號展頻碼；

一領航信號展頻碼匹配濾波器，用以從該領航干擾消除單元中該估計得到的該領航信號成分中去除領航信號展頻碼，並產生通道頻率響應估計；

一通道匹配濾波器，用以產生該資訊信號展頻碼匹配濾波器之輸出與該通道頻率響應估計之共軛複數的乘積，最後計算該乘積在一展頻碼週期之積項和；以及

一判定單元，基於一預選調變形式以從該通道匹配濾波器產生之積項和來判定該資訊值。

20.如申請專利範圍第11項所述之該下鏈接收機，其中該領航干擾消除單元包括：

一領航信號估計單元，用以從該接收基頻信號之快速傅利葉轉換中估計該領航信號成分；以及

一結合器，用以從該接收基頻信號之快速傅利葉轉換中減去該估計領航信號成分。

21.如申請專利範圍第20項所述之該下鏈接收機，其中該領航信號估計單元包括：

一延遲單元，延遲該接收基頻信號之快速傅利葉轉換一展頻碼週期；以及

一平均單元，執行該延遲後接收基頻信號之快速傅利葉轉換之加權平均。

22.如申請專利範圍第19項所述之該下鏈接收機，其中該領航信號估計單元包括：

一延遲單元，延遲該接收基頻信號之快速傅利葉轉換一展頻碼週期；

一資訊信號重建單元，以一判定回授的方法為基礎重建該資訊信號成分；

一領航信號估計結合器，從該延遲後接收基頻信號之快速傅利葉轉換中減去該重建資訊信號成分；以及

一平均單元，用以執行該領航信號估計結合器其結果之一加權平均。

23.如申請專利範圍第19項所述之該下鏈接收機，其中該通道匹配濾波器包括：

一共軛複數單元，用以計算該通道頻率響應估計之共軛複數；

一乘法器，將該資訊信號展頻碼匹配濾波器之輸出與該共軛複數單元之輸出相乘；以及

一加法器，執行該乘法器其乘積在一展頻碼週期內之一積項和。

24.如申請專利範圍第19項所述之該下鏈接收機，其中該通道匹配濾波器還包括一保留主要路徑單元以保留複數主要路徑，該保留主路徑單元包括：

一反快速傅利葉反轉換(IFFT)單元，用以計算該通道頻率響應估計之反快速傅利葉轉換以產生一通道脈衝響應估計；

一搜尋單元，用以從該通道脈衝響應估計中搜尋振幅峰值；

一選取單元，以該搜尋得之振幅峰值為基礎產生一保留主要路徑之通道脈衝響應估計；以及

一快速傅利葉轉換單元，用以計算該保留主要路徑之通道脈衝響應估計之快速傅利葉轉換。

25.一種上鏈接收機，用以接收符號之資

訊檢測，其包括：

一多人干擾消除單元，用以從一接收基頻信號之快速傅利葉轉換中減去估計之複數多重進接干擾信號；

一資訊信號展頻碼匹配濾波器，用以從該多人干擾消除單元輸出中去除資訊信號展頻碼；

一領航信號展頻碼匹配濾波器，用以從該多人干擾消除單元之輸出中去除領航信號展頻碼；

一開關，連接於該多人干擾消除單元與該領航信號展頻碼濾波器之間，當該本接收符號之該最後一級資訊檢測到達時關閉；

一通道匹配濾波器，以該領航信號展頻碼匹配濾波器之輸出為基礎產生一通道頻率響應估計，然後產生該資訊信號展頻碼匹配濾波器之輸出與該通道頻率響應估計之共軛複數之乘積，最後計算該乘積於一展頻碼週期內之積項和；

一判定單元，用以基於一預選調變形式為基礎，以該通道匹配濾波器產生之積項和判定一實驗資訊值，在最後一級資訊檢測後，該實驗資訊值為本接收符號之最終資訊判定；以及

一干擾信號估計單元，利用基於該通道頻率響應估計、該領航信號展頻碼之快速傅利葉轉換、資訊信號展頻碼之快速傅利葉轉換、與該判定之實驗資訊值來估計多重進接干擾。

26.如申請專利範圍第25項之該上鏈接收機，其中該多人干擾消除單元包括：

一第一加法器，用以將來自其他複數接收機中該等估計多重進接干擾信號相加；以及

一第二加法器，用以從該接收基頻信號之快速傅利葉轉換中將該等相加之估計多重進接干擾信號減去。

·27.如申請專利範圍第25項之該上鏈接收

機，其中該資訊信號展頻碼匹配濾波器包括：

一儲存單元，用以儲存一資訊信號展頻碼之快速傅利葉轉換；

一共軛複數單元，用以計算該資訊信號展頻碼其快速傅利葉轉換之共軛複數；以及

一乘法器，用以將該多人干擾消除單元之輸出與該共軛複數單元之輸出相乘。

28.如申請專利範圍第25項之該上鏈接收機，其中該領航信號展頻碼匹配濾波器包括：

一儲存單元，用以儲存一領航信號展頻碼之快速傅利葉轉換；

一共軛複數單元，用以計算該領航信號展頻碼其快速傅利葉轉換之共軛複數；以及

一乘法器，用以在該最後一級資訊檢測時將多人干擾消除單元之輸出與共軛複數單元之輸出相乘。

29.如申請專利範圍第25項之該上鏈接收機，其中該通道匹配濾波器包括該通道頻率響應估計單元，以該領航信號展頻碼匹配濾波器之該輸出為基礎產生一通道頻率響應估計，該通道頻率響應估計單元包括：

一延遲單元，延遲該領航信號展頻碼匹配濾波器之輸出直到下一個接收符號之資訊檢測；

一平均單元，用以執行該延遲後領航信號展頻碼匹配濾波器輸出之一加權平均；以及

一共軛複數單元，用以計算該延遲後領航信號展頻碼輸出之該加權平均之共軛複數。

30.如申請專利範圍第29項所述之該上鏈接收機，其中該通道頻率響應估計單元包括一保留主要路徑單元用以保留複數主要路徑，該保留主要路徑單元

包括：

一反快速傅利葉轉換單元(IFFT)，計算該延遲後領航信號展頻碼匹配濾波器輸出之加權平均之反快速傅利葉轉換，而產生一通道脈衝響應估計；

一搜尋單元，從該通道脈衝響應估計中搜尋出振幅峰值；

一選取單元，以該搜尋得之振幅峰值為基礎，選取複數主要路徑，然後產生一保留主要路徑之通道脈衝響應估計；以及

一快速傅利葉轉換單元，用以計算出該保留主要路徑之通道脈衝響應估計之快速傅利葉轉換。

31.如申請專利範圍第29項所述之該上鏈接收機，其中該通道匹配濾波器還包括：

一乘法器，用以將該資訊信號展頻碼匹配濾波器之輸出與該通道頻率響應估計之共軛複數相乘；以及

一加法器，用以計算在一展頻碼週期中該乘法器其輸出之一積項和。

32.如申請專利範圍第25項所述之該上鏈接收機，其中該干擾信號估計單元包括：

一第一乘法器，用以將該資訊信號展頻碼信號之快速傅利葉轉換與一該實驗資訊判定相乘；

一加法器，用以將該第一乘法器之輸出與該領航信號展頻碼之快速傅利葉轉換相加；

一第二乘法器，用以將該通道頻率響應估計與該加法器之輸出相乘；以及

一正規化單元，以該領航信號展頻碼其快速傅利葉轉換之平方模正規化該第二乘法器之輸出。

圖式簡單說明：

第一圖A與第一圖B係分別顯示作為下鏈與上鏈犁耙式接收機之一系統架構圖；

第二圖A係顯示一下鏈犁耙式接收機之方塊圖包含選擇性地執行保留複數主路徑與資訊信號重建；

第二圖B係顯示另一種可替換之(alternate)下鏈犁耙式接收機之方塊圖；

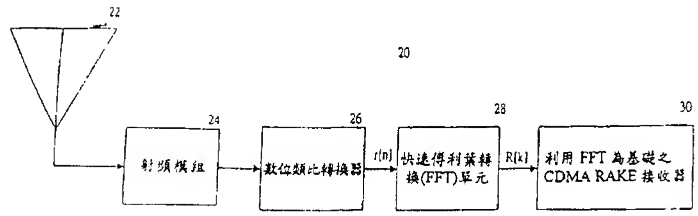
第三圖A係顯示一上鏈犁耙式接收機之方塊圖包含選擇性地執行保留複數主路徑；

第三圖B係顯示第三圖A中該上鏈犁耙式接收機之資訊檢測流程圖；

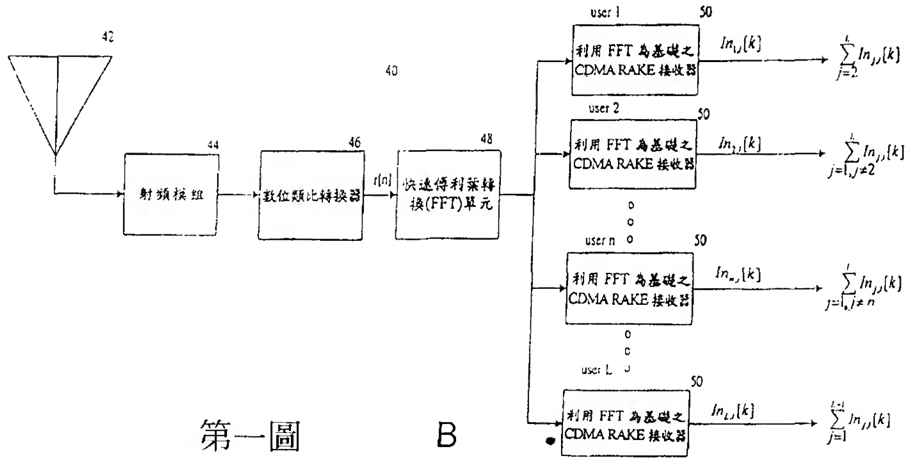
第四圖A與第四圖B係顯示第二圖A與第二圖B中該等下鏈犁耙式接收機兩種執行領航信號估計之方塊圖；

第五圖係顯示第二圖A與第二圖B中該等下鏈犁耙式接收機其選擇性地執行資訊信號重建之方塊圖；以及

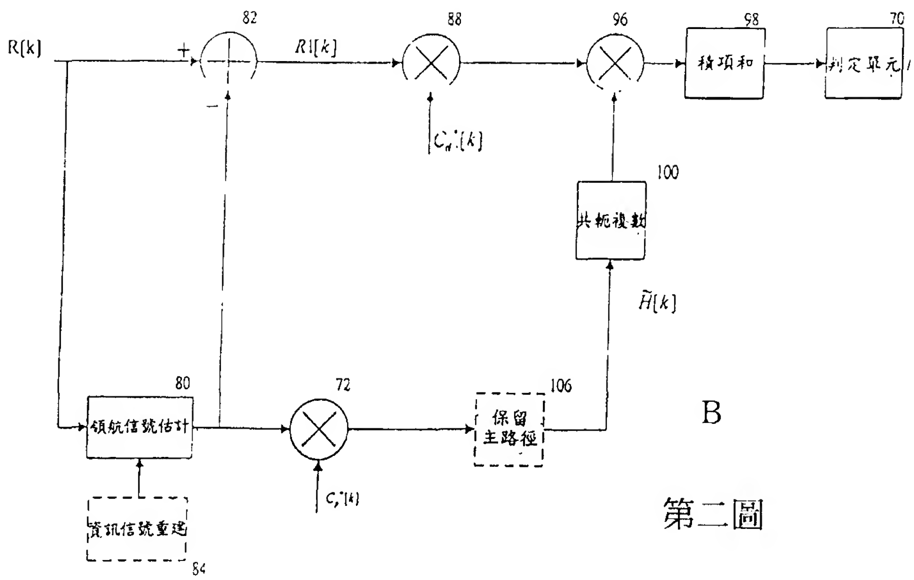
第六圖係顯示第二圖A、第二圖B與第三圖A中該等犁耙式接收機其選擇性地執行保留複數主路徑之方塊圖。


22
20
24
26
28
30
射頻模組
數位類比轉換器
r[n]
快速傅利葉轉換(FFT)單元
R[k]
利用 FFT 為基礎之 CDMA RAKE 接收器

A

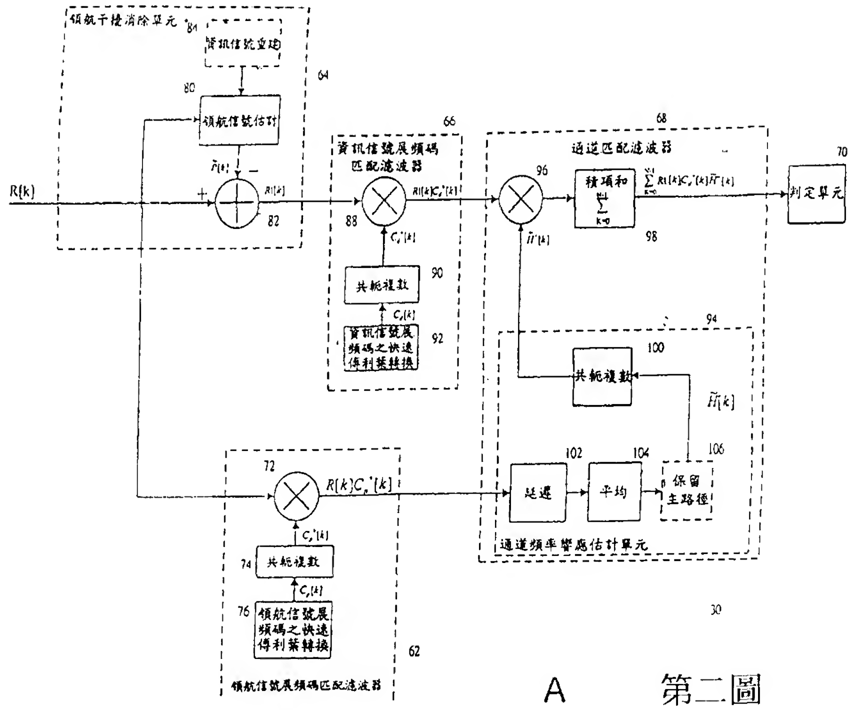

42
40
44
46
48
50
射頻模組
數位類比轉換器
r[n]
快速傅利葉轉換(FFT)單元
user 1
user 2
user n
user L
利用 FFT 為基礎之 CDMA RAKE 接收器

第一圖 B


110
R[k]
82
88
96
98
70
積項和
判定單元
100
共軛複數
80
領航信號估計
72
106
保留主路徑
資訊信號重建
84

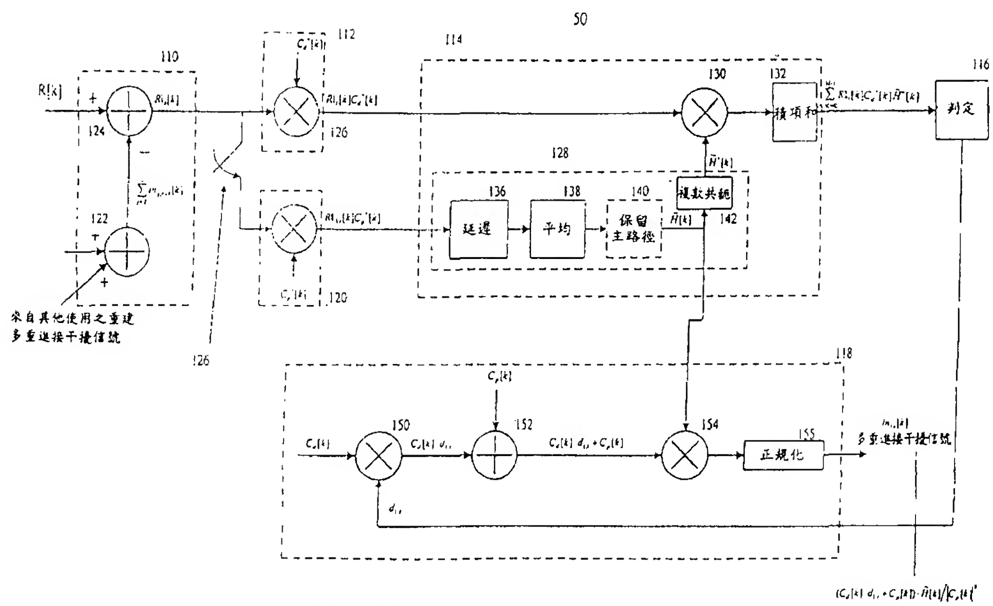
B

第二圖



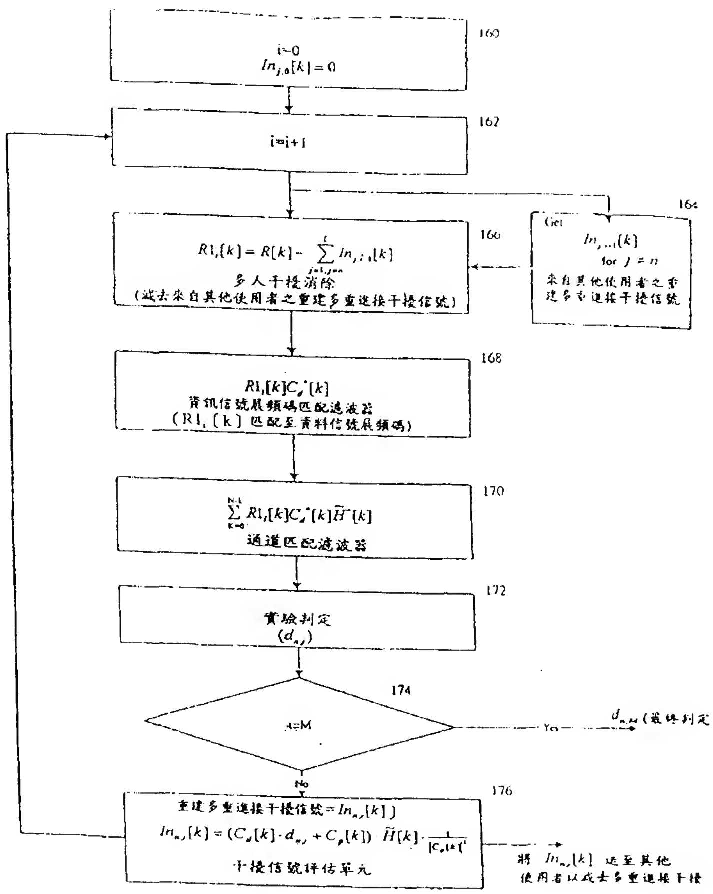
領航干擾消除單元
資訊信號重建
領航信號估計
R(k)
資訊信號展頻碼匹配濾波器
共軛複數
資訊信號展頻碼之快速傅利葉轉換
通道匹配濾波器
積項和
判定單元
共軛複數
延遲
平均
保留主路徑
通道頻率響應估計單元
共軛複數
領航信號展頻碼之快速傅利葉轉換
領航信號展頻碼匹配濾波器

A 第二圖



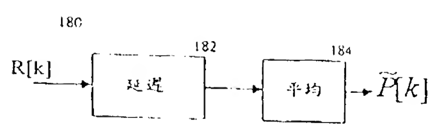
R[k]
積項和
判定
延遲
平均
保留主路徑
複數共軛
來自其他使用之重建多重進接干擾信號
正規化
多重進接干擾信號

第三圖 A

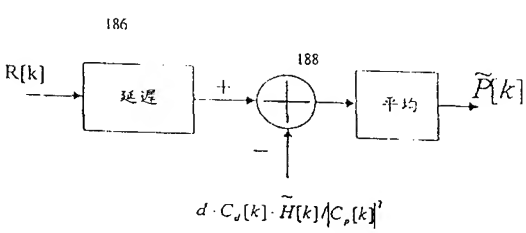

160
i=0
$In_{j,0}[k]=0$
162
i=i+1
164
Get
$In_{j,i-1}[k]$
for $j\neq n$
來自其他使用者之重建多重進接干擾信號
166
$R1_i[k]=R[k]-\sum_{j=1,j\neq n}^{L} In_{j,i-1}[k]$
多人干擾消除
(減去來自其他使用者之重建多重進接干擾信號)
168
$R1_i[k]C_d^*[k]$
資訊信號展頻碼匹配濾波器
(R1$_i$[k]匹配至資料信號展頻碼)
170
$\sum_{k=0}^{N-1} R1_i[k]C_d^*[k]\widetilde{H}^*[k]$
通道匹配濾波器
172
實驗判定
$(d_{n,i})$
174
i=M
Yes
$d_{n,M}$(最終判定)
No
176
重建多重進接干擾信號$=In_{n,i}[k]$
干擾信號評估單元
將$In_{n,i}[k]$送至其他使用者以減去多重進接干擾

B

第三圖

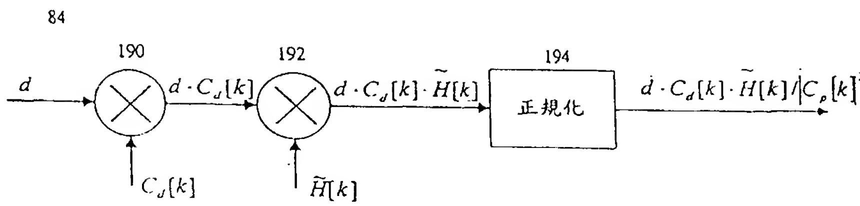

180
182
184
R[k]
延遲
平均
$\widetilde{P}[k]$

A

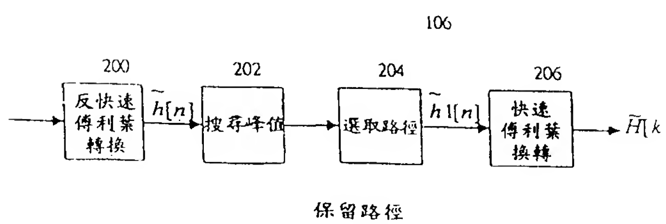

186
188
R[k]
延遲
+
−
平均
$\widetilde{P}[k]$
$d\cdot C_d[k]\cdot\widetilde{H}[k]/|C_p[k]|^2$

B

第四圖


84
190
192
194
d
$d \cdot C_d[k]$
$d \cdot C_d[k] \cdot \tilde{H}[k]$
正規化
$d \cdot C_d[k] \cdot \tilde{H}[k] / |C_p[k]|^2$
$C_d[k]$
$\tilde{H}[k]$

第五圖


106
200
202
204
206
反快速傅利葉轉換
$\tilde{h}[n]$
搜存峰值
選取路徑
$\tilde{h}1[n]$
快速傅利葉換轉
$\tilde{H}[k]$
保留路徑

第六圖

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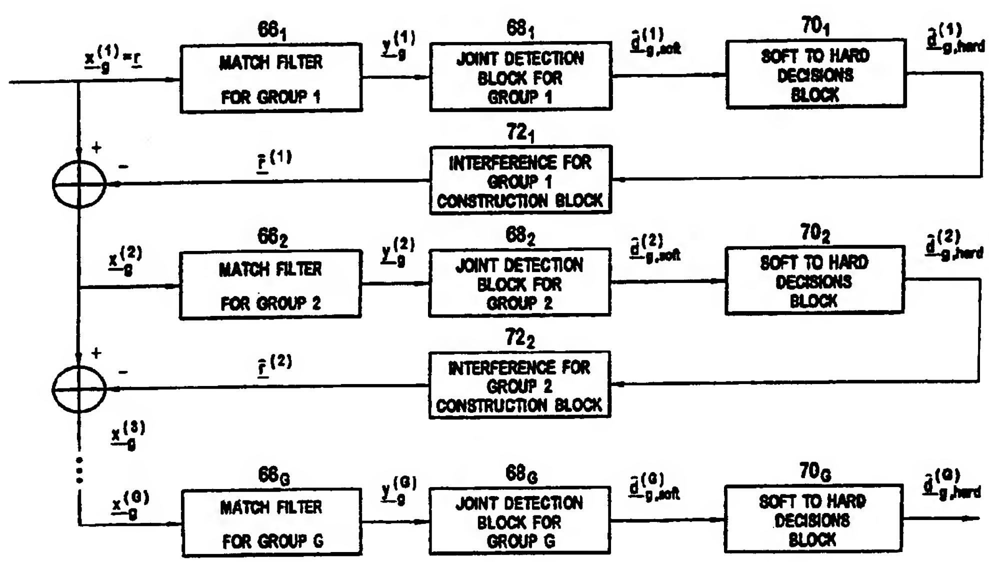

(57) Abstract: A time division duplex communication system using code division multiple access transmits a plurality of data signals over a shared spectrum in a time slot. A combined signal is received over the shared spectrum in the time slot. The plurality of data signals are grouped into a plurality of groups. The combined signal is matched filtered based on in part symbol responses associated with the data signals of one of the groups. Data from each data signal in the one group is jointly detected. An interference signal is constructed based on in part the one group detected data. The constructed interference signal is subtracted from the combined signal. Data from the other groups is detected by processing the subtracted signal.

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MULTI-USER DETECTION USING AN ADAPTIVE COMBINATION OF JOINT DETECTION AND SUCCESSIVE INTERFERENCE CANCELLATION

This application claims priority to U.S. Provisional Patent Application No. 60/189,680, filed on March 15, 2000 and U.S. Provisional Patent Application No. 60/207,700, filed on May 26, 2000.

BACKGROUND

The invention generally relates to wireless communication systems. In particular, the invention relates to joint detection of multiple user signals in a wireless communication system.

Figure 1 is an illustration of a wireless communication system 10. The communication system 10 has base stations 12_1 to 12_5 which communicate with user equipments (UEs) 14_1 to 14_3. Each base station 12_1 has an associated operational area where it communicates with UEs 14_1 to 14_3 in its operational area.

In some communication systems, such as code division multiple access (CDMA) and time division duplex using code division multiple access (TDD/CDMA), multiple communications are sent over the same frequency spectrum. These communications are typically differentiated by their chip code sequences. To more efficiently use the frequency spectrum, TDD/CDMA communication systems use repeating frames divided into time slots for communication. A communication sent in such a system will have one or multiple associated chip codes and time slots assigned to it based on the communication's bandwidth.

Since multiple communications may be sent in the same frequency spectrum and at the same time, a receiver in such a system must distinguish between the multiple communications. One approach to detecting such signals is matched

filtering. In matched filtering, a communication sent with a single code is detected. Other communications are treated as interference. To detect multiple codes, a respective number of matched filters are used. Another approach is successive interference cancellation (SIC). In SIC, one communication is detected and the contribution of that communication is subtracted from the received signal for use in detecting the next communication.

In some situations, it is desirable to be able to detect multiple communications simultaneously in order to improve performance. Detecting multiple communications simultaneously is referred to as joint detection. Some joint detectors use Cholesky decomposition to perform a minimum mean square error (MMSE) detection and zero-forcing block equalizers (ZF-BLEs). These detectors have a high complexity requiring extensive receiver resources.

Accordingly, it is desirable to have alternate approaches to multi-user detection.

SUMMARY

A time division duplex communication system using code division multiple access transmits a plurality of data signals over a shared spectrum in a time slot. A combined signal is received over the shared spectrum in the time slot. The plurality of data signals are grouped into a plurality of groups. The combined signal is matched filtered based on in part symbol responses associated with the data signals of one of the groups. Data from each data signal in the one group is jointly detected. An interference signal is constructed based on in part the one group detected data. The constructed interference signal is subtracted from the combined signal. Data from the other groups is detected by processing the subtracted signal.

BRIEF DESCRIPTION OF THE DRAWING(S)

Figure 1 is a wireless communication system.

Figure 2 is a simplified transmitter and a receiver using joint detection.

Figure 3 is an illustration of a communication burst.

Figure 4 is a flow chart of adaptive combination of joint detection and successive interference cancellation.

Figure 5 is an illustration of an adaptive combination of joint detection and successive interference cancellation device.

Figures 6-12 are graphs comparing the performance of adaptive combination of joint detection and successive interference cancellation, full joint detection and a RAKE receiver.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT(S)

Figure 2 illustrates a simplified transmitter 26 and receiver 28 using an adaptive combination of joint detection (JD) and successive interference cancellation (SIC), "SIC-JD", in a TDD/CDMA communication system. In a typical system, a transmitter 26 is in each UE 14_1 to 14_3 and multiple transmitting circuits 26 sending multiple communications are in each base station 12_1 to 12_5. A base station 12_1 will typically require at least one transmitting circuit 26 for each actively communicating UE 14_1 to 14_3. The SIC-JD receiver 28 may be at a base station 12_1, UEs 14_1 to 14_3 or both. The SIC-JD receiver 28 receives communications from multiple transmitters 26 or transmitting circuits 26.

Each transmitter 26 sends data over a wireless radio channel 30. A data generator 32 in the transmitter 26 generates data to be communicated over a reference channel to a receiver 28. Reference data is assigned to one or multiple codes and/or time slots based on the communications bandwidth requirements. A modulation and spreading device 34 spreads the reference data and makes the spread reference data time-multiplexed with a training sequence in the appropriate assigned time slots and codes. The resulting sequence is referred to as a communication burst. The communication burst is modulated by a modulator 36 to radio frequency. An antenna 38 radiates the RF signal through the wireless radio channel 30 to an antenna 40 of the receiver 28. The type of modulation used for the transmitted

communication can be any of those known to those skilled in the art, such as direct phase shift keying (DPSK) or quadrature phase shift keying (QPSK).

A typical communication burst 16 has a midamble 20, a guard period 18 and two data bursts 22, 24, as shown in Figure 3. The midamble 20 separates the two data bursts 22, 24 and the guard period 18 separates the communication bursts to allow for the difference in arrival times of bursts transmitted from different transmitters. The two data bursts 22, 24 contain the communication burst's data and are typically the same symbol length. The midamble contains a training sequence.

The antenna 40 of the receiver 28 receives various radio frequency signals. The received signals are demodulated by a demodulator 42 to produce a baseband signal. The baseband signal is processed, such as by a channel estimation device 44 and a SIC-JD device 46, in the time slots and with the appropriate codes assigned to the communication bursts of the corresponding transmitters 26. The channel estimation device 44 uses the training sequence component in the baseband signal to provide channel information, such as channel impulse responses. The channel information is used by the SIC-JD device 46 to estimate the transmitted data of the received communication bursts as hard symbols.

The SIC-JD device 46 uses the channel information provided by the channel estimation device 44 and the known spreading codes used by the transmitters 26 to estimate the data of the various received communication bursts. Although SIC-JD is described in conjunction with a TDD/CDMA communication system, the same approach is applicable to other communication systems, such as CDMA.

One approach to SIC-JD in a particular time slot in a TDD/CDMA communication system is illustrated in Figure 4. A number of communication bursts are superimposed on each other in the particular time slot, such as K communication bursts. The K bursts may be from K different transmitters. If certain transmitters are using multiple codes in the particular time slot, the K bursts may be from less than K transmitters.

Each data burst 22, 24 of the communication burst 16 has a predefined number of transmitted symbols, such as N_S. Each symbol is transmitted using a predetermined number of chips of the spreading code, which is the spreading factor (SF). In a typical TDD communication system, each base station 12_1 to 12_5 has an associated scrambling code mixed with its communicated data. The scrambling code distinguishes the base stations from one another. Typically, the scrambling code does not affect the spreading factor. Although the terms spreading code and factor are used hereafter, for systems using scrambling codes, the spreading code for the following is the combined scrambling and spreading codes. As a result, each data burst 22, 24 has N_S x SF chips. After passing through a channel having an impulse response of W chips, each received burst has a length of SF x N_s + W -1, which is also represented as N_c chips. The code for a k^{th} burst of the K bursts is represented by $C^{(k)}$.

Each k^{th} burst is received at the receiver and can be represented by Equation 1.

$$\underline{r}^{(k)} = A^{(k)}\underline{d}^{(k)},\ k = 1 \cdots K \qquad \text{Equation 1}$$

$\underline{r}^{(k)}$ is the received contribution of the k^{th} burst. $A^{(k)}$ is the combined channel response, being an N_c x N_s matrix. Each j^{th} column in $A^{(k)}$ is a zero-padded version of the symbol response $s^{(k)}$ of the j^{th} element of $d^{(k)}$. The symbol response $s^{(k)}$ is the convolution of the estimated response $\underline{h}^{(k)}$ and spreading code $C^{(k)}$ for the burst. $\underline{d}^{(k)}$ is the unknown data symbols transmitted in the burst. The estimated response for each k^{th} burst, $\underline{h}^{(k)}$, has a length W chips and can be represent by Equation 2.

$$\underline{h}^{(k)} = \gamma^{(k)} \cdot \underline{\tilde{h}}^{(k)} \qquad \text{Equation 2}$$

$\gamma^{(k)}$ reflects the transmitter gain and/or path loss. $\underline{\tilde{h}}^{(k)}$ represents the burst-specific fading channel response or for a group of bursts experiencing a similarly channel, $\underline{\tilde{h}}^{(g)}$ represents the group-specific channel response. For uplink communications, each $\underline{h}^{(k)}$ as well as each $\gamma^{(k)}$ and $\underline{\tilde{h}}^{(k)}$ are distinct. For the downlink, all of the bursts have the same $\underline{\tilde{h}}^{(k)}$ but each $\gamma^{(k)}$ is different. If transmit diversity is used in the downlink, each $\gamma^{(k)}$ and $\underline{\tilde{h}}^{(k)}$ are distinct.

The overall received vector from all K bursts sent over the wireless channel is per Equation 3.

$$\underline{r} = \sum_{i=1}^{K} \underline{r}^{(k)} + \underline{n}$$

Equation 3

$\underline{n}$ is a zero-mean noise vector.

By combining the $A^{(k)}$ for all data bursts into matrix A and all the unknown data for each burst $\underline{d}^{(k)}$ into matrix $\underline{d}$, Equation 1 becomes Equation 4.

$$\underline{r} = A\underline{d} + \underline{n}$$

Equation 4

SIC-JD determines the received power of each k^{th} burst. This determination may be based on apriori knowledge at the receiver 28, burst-specific channel estimation from a burst-specific training sequence, or a bank of matched filters. The K bursts are arranged in descending order based on the determined received power.

Bursts having roughly the same power level, such as within a certain threshold, are grouped together and are arranged into G groups, 48. The G groups are arranged into descending order by their power, such as from group 1 to G with

group 1 having the highest received power. Figure 5 is an illustration of a SIC-JD device 46 performing SIC-JD based on the G groups.

For the group with the highest received power, group 1, the symbol response matrix for only the bursts in group 1, $A_g^{(1)}$, is determined. $A_g^{(1)}$ contains only the symbol responses of the bursts in group 1. The received vector, $\underline{r}$, is modeled for group 1 as $\underline{x}_g^{(1)}$. As a result, Equation 4 becomes Equation 5 for group 1.

$$\underline{x}_g^{(1)} = A_g^{(1)} \underline{d}_g^{(1)} + \underline{n}$$

Equation 5

$\underline{d}_g^{(1)}$ is the data in the bursts of group 1. Equation 5 addresses both the effects of inter symbol interference (ISI) and multiple access interference (MAI). As a result, the effects of the other groups, groups 2 to G, are ignored.

The received vector, $\underline{x}_g^{(1)}$, is matched filtered to the symbol responses of the bursts in group 1 by a group 1 matched filter 66_1, such as per Equation 6, 50.

$$\underline{y}_g^{(1)} = A_g^{(1)^H} \underline{x}_g^{(1)}$$

Equation 6

$\underline{y}_g^{(1)}$ is the matched filtered result.

A joint detection is performed on group 1 by a group 1 joint detection device 68_1 to make a soft decision estimate of $\hat{d}_{g,soft}^{(1)}$, using the matched filtered result $\underline{y}_g^{(1)}$. One JD approach is to compute the least-squares, zero-forcing, solution of Equation 7.

$$\hat{\underline{d}}_{g,soft}^{(1)} = \left(A_g^{(1)^H} A_g^{(1)}\right)^{-1} \underline{y}_g^{(1)}$$

Equation 7

$A_g^{(1)^H}$ is the hermetian of $A_g^{(1)}$. Another JD approach is to compute the minimum mean square error solution (MMSE) as per Equation 8.

$$\hat{\underline{d}}^{(1)}_{g,soft} = \left(A_g^{(1)^H} A_g^{(1)} + \sigma^2 I\right)^{-1} \underline{y}_g^{(1)}$$

Equation 8

I is the Identity matrix and σ^2 is the standard deviation.

One advantage to performing joint detection on only a group of bursts is that the complexity of analyzing a single group versus all the signals is reduced. Since $A_g^{(1)^H}$ and $A_g^{(1)}$ are banded block Toeplitz matrices, the complexity in solving either Equation 7 or 8 is reduced. Additionally, Cholesky decomposition may be employed with a negligible loss in performance. Cholesky decomposition performed on a large number of bursts is extremely complex. However, on a smaller group of users, Cholesky decomposition can be performed at a more reasonable complexity.

The soft decisions, $\hat{d}^{(1)}_{g,soft}$, are converted into hard decisions, $\hat{d}^{(1)}_{g,hard}$, by soft to hard decision block 70_1 as the received data for group 1, 54. To process the other weaker groups, the multiple access interference caused by group 1 onto the weaker groups is estimated by a group 1 interference construction block 72_1 using Equation 9, 56.

$$\hat{\underline{r}}^{(1)} = A_g^{(1)} \hat{\underline{d}}^{(1)}_{g,hard}$$

Equation 9

$\hat{\underline{r}}^{(1)}$ is the estimated contribution of group 1 to $\underline{r}$.

For the next group 2, the estimated contribution of group 1 is removed from the received vector, $\underline{x}_g^{(1)}$, to produce $\underline{x}_g^{(2)}$, such as by a subtractor 74_1, as per Equation 10, 58.

$$\underline{x}_g^{(2)} = \underline{x}_g^{(1)} - \hat{\underline{r}}^{(1)}$$

Equation 10

As a result, multiple access interference from group 1 is effectively canceled from the received signal. The next strongest group, group 2, is processed similarly using $\underline{x}_g^{(2)}$, with group 2 matched filter 66_2, group 2 JD block 68_2, soft to hard decision block 70_2 and group 2 interference construction block 72_2, 60. The constructed group 2 interference, $\underline{\hat{r}}^{(2)}$, is subtracted, such as by subtractor 24_2, from the interference cancclled signal for group 2, $\underline{x}_g^{(2)} - \underline{\hat{r}}^{(2)} = \underline{x}_g^{(3)}$, 62. Using this procedure, each group is successively processed until the final group G. Since group G is the last group, the interference construction does not need to be performed. Accordingly, group G is only processed with group G matched filter 66_G, group G JD block 68_G and soft to hard decisions block 70_G to recovery the hard symbols, 64.

When SIC-JD is performed at a UE 14_1, it may not be necessary to process all of the groups. If all of the bursts that the UE 14_1 is intended to receive are in the highest received power group or in higher received power groups, the UE 14_1 will only have to process the groups having its bursts. As a result, the processing required at the UE 14_1 can be further reduced. Reduced processing at the UE 14_1 results in reduced power consumption and extended battery life.

SIC-JD is less complex than a single-step JD due to the dimension N_c x $K \cdot N_s$ matrix being replaced with G JD stages of dimension N_c x $n_i \cdot N_s$, where i = 1 to G. n_i is the number of bursts in the i^{th} group. The complexity of JD is proportional to the square to cube of the number of bursts being jointly detected.

An advantage of this approach is that a trade-off between computational complexity and performance can be achieved. If all of the bursts are placed in a single group, the solution reduces to a JD problem. The single grouping can be achieved by either forcing all the bursts into one group or using a broad threshold. Alternately, if the groups contain only one signal or only one signal is received, the solution reduces to a SIC-LSE problem. Such a situation could result using a narrow threshold or forcing each burst into its own group, by hard limiting the group size.

By selecting the thresholds, an optional tradeoff between performance and complexity can be achieved.

Figures 6 to 12 are simulation results that compare the bit error rate (BER) performance of SIC-JD to full JD and RAKE-like receivers under various multi-path fading channel conditions. The parameters chosen are those of the 3G UTRA TDD CDMA system: SF = 61 and W = 57. Each TDD burst/time-slot is 2560 chips or 667 microseconds long. The bursts carry two data fields with N_S QPSK symbols each, a midamble field and a guard period. Each simulation is run over 1000 timeslots. In all cases the number of bursts, K is chosen to be 8. All receivers are assumed to have exact knowledge of the channel response of each burst, which is used to perfectly rank and group the bursts. The channel response is assumed to be time-invariant over a time-slot, but successive time-slots experience uncorrelated channel responses. No channel coding was applied in the simulation. The JD algorithm jointly detects all K bursts. The RAKE-like receiver was a bank of matched filters, $\underline{\hat{d}}^{(i)} = A^{(i)^H}\underline{r}^{(i)}$, for an i^{th} burst's code. The maximal ratio combiner (MRC) stage is implicit in these filters because they are matched to the entire symbol-response.

The performance was simulated under fading channels with multi-path profiles defined by the ITU channel models, such as the Indoor A, Pedestrian A, Vehicular A models, and the 3GPP UTRA TDD Working Group 4 Case 1, Case 2 and Case 3 models. In Vehicular A and Case 2 channels, the SIC-JD suffered a degradation of up to 1 decibel (dB) as compared to the full JD in the 1% to 10% BER range. For all other channels, the SIC-JD performance was within 0.5 dB of that of the full JD. Since Vehicular A and Case 2 represent the worst-case amongst all cases studied, only the performance plots are shown. Amongst all channels simulated, Vehicular A and Case 2 have the largest delay spread. Vehicular A is a six tap model with relative delays of 0, 310, 710, 1090, 1730 and 2510 nanoseconds and relative average powers of 0, -1, -9, -10, -15 and -20 decibels (dB). Case 2 is a

3 tap model, all with the same average power and with relative delays of 0, 976 and 1200 nanoseconds.

Figures 6 and 7 compare the bit error rate (BER) vs. the chip-level signal to noise ratio (SNR) performance of the SIC-LSE receiver with the full JD and RAKE-like receivers under two multi-path fading channel conditions. The group size is forced to be 1, to form K groups, both, at the transmitter and receiver. The theoretical binary phase shift keying (BPSK) BER in an additive white gaussian noise (AWGN) channel that provides a lower bound to the BER is also shown. The BER is averaged over all bursts. Figure 6 represents the distinct channel case wherein each burst is assumed to pass through an independently fading channel but all channels have the same average power leading to the same average SNR. Thus, in this case, $\underline{\tilde{h}}^{(i)}, i=1\cdots K$ are distinct while $\gamma^{(i)}, i=1\cdots K$ are all equal. Such a situation exists in the uplink where the power control compensates for long-term fading and/or path-loss but not for short-term fading. At each time-slot, the bursts were arranged in power based upon the associated $\underline{h}^{(i)}, i=1\cdots K$. Figure 7 shows similar plots for the common channel case. All bursts are assumed to pass through the same multi-path channel, i.e., $\underline{\tilde{h}}^{(i)}, i=1\cdots K$ and are all equal, but with different $\gamma^{(i)}, i=1\cdots K$. The $\delta^{(i)}$ are chosen such that neighboring bursts have a power separation of 2 dB when arranged by power level. Such difference in power can exist, for instance, in the downlink where the base station 12_1 applies different transmit gains to bursts targeted for different UEs 14_1 to 14_3. Figures 6 and 7 show that in the range of 1% to 10% bit error rate (BER), the SIC-LSE suffers a degradation of less than 1 dB as compared to the JD. This is often the range of interest for the uncoded BER (raw BER). The RAKE receiver exhibits significant degradation, since it does not optimally handle the ISI. As the power differential between bursts increases, the performance of SIC-LSE improves. Depending upon

the channel, a power separation of 1 to 2 dB is sufficient to achieve a performance comparable to that of the full JD.

Figures 8, 9, 10 and 11 compare the BER vs. SNR performance of the SIC-JD receiver with the full JD and RAKE-like receivers under two multi-path fading channels. The 8 codes are divided into 4 groups of 2 codes each at the transmitter and receiver. The BER is averaged over all bursts. Figures 8 and 9 represent the distinct channel case wherein different groups are assumed to pass through independently fading channels. However, all channels have the same average power leading to the same average SNR. All bursts within the same group are subjected to an identical channel response. In this case, $\underline{\tilde{h}}_g^{(g)}, g = 1 \cdots G$ are all distinct, but the channel responses, $h_g^{(i)}, i = 1, \cdots n_g$, for each burst in the group are equal. n_g is the number of bursts in the g^{th} group. This potentially represents a multi-code scenario on the uplink, where each UE 14_1 transmits two codes. The SIC-JD receiver 28 groups the multi-codes associated with a single UE 14_1 into the same group, thus forming 4 groups. Figures 10 and 11 represent the common channel case. All groups are assumed to pass through the same multi-path channel, i.e., $\underline{\tilde{h}}_g^{(i)}, g = 1 \cdots n_g$ are all equal, but with different $\gamma_g, g = 1 \cdots G$. The γ_g are chosen such that, when arranged according to power, neighboring groups have a power separation of 2 dB. This potentially represents a multi-code scenario on the downlink where the base station 12_1 transmits 2 codes per UE 14_1. Figures 10 and 11 show a trend similar to that observed for the SIC-LSE shown in Figures 8 and 9. SIC-JD has a performance comparable (within a dB) to the JD in the region of 1% to 10% BER, which is the operating region of interest for the uncoded BER. Depending upon the channel, a power separation of 1 to 2 dB is sufficient to achieve a performance of SIC-LSE comparable to that of the full JD. As shown, performance improves as the power separation between bursts increases.

Figure 12 is similar to Figure 10, except that there are only two groups with 4 bursts each. As shown in Figure 12, SIC-JD has a performance comparable (within a dB) to JD in the region of 1% to 10% BER.

The complexity of SIC-JD is less than full JD. The reduced complexity stems from the replacement of a single-step JD which is a dimension $N_c \times K \cdot N_s$ with G JD stages of dimension $N_c \times n_i \cdot N_s$, $i = 1 \cdots G$. Since, typically, JD involves a matrix inversion, whose complexity varies as the cube of the number of bursts, the overall complexity of the multi-stage JD can be significantly lower than that of the single-stage full JD. Furthermore, the complexity of the SIC part varies only linearly with the number of bursts, hence it does not offset this complexity advantage significantly. For instance, the complexity of the G-1 stages of interference cancellation can be derived as follows. Since successive column blocks of $A_g^{(i)}$ are shifted versions of the first block and assuming that elements of $\underline{\hat{d}}_{g,hard}^{(i)}$ belong to 1 of 4 QPSK constellation points, the $4 \cdot n_i$ possible vectors can be computed that are needed in computing the product $A_g^{(i)} \underline{\hat{d}}_{g,hard}^{(i)}$. This step requires $4\alpha \cdot (SF + W - 1) \cdot \frac{Rate}{10^6} \sum_{i=1}^{G-1} n_i$ million real operations per sec (MROPS). $\alpha = 4$ is the number of real operations per complex multiplication or multiply and accumulate (MAC). *Rate* is the number of times the SIC-JD is performed per second. With these $4 \cdot n_i$ vectors already computed, the computation of $\underline{x}_g^{(i+1)}$ requires $\frac{\alpha}{2} \cdot N_s \cdot (SF + W - 1) \cdot \frac{Rate}{10^6} \sum_{i=1}^{G-1} n_i$ MROPS. The factor of $\frac{\alpha}{2}$ comes from the fact that only complex additions are involved. Hence, only 2 real operations are required for

each complex operation. It then follows that the complexity of $G-1$ stages of interference cancellation can be expressed by Equation 11.

$$Z = \alpha(SF + W - 1) \cdot \left(4 + \frac{N_s}{2}\right) \cdot \frac{Rate}{10^6} \cdot \sum_{i=1}^{G-1} n_i$$

Equation 11

The complexity of converting soft to hard decisions is negligible.

There are several well-known techniques to solve the matrix inversion of JD. To illustrate the complexity, an approach using a very efficient approximate Cholesky factor algorithm with negligible loss in performance as compared to the exact Cholesky factor algorithm was used. The same algorithm can be employed to solve group-wise JD. The complexity of the full JD and the SIC-JD for the 3GPP UTRA TDD system is shown in Table 1. Table 1 compares their complexity for various group sizes. It can be seen that as K increases or as the group size decreases the complexity advantage of the SIC-JD over the full JD increases. The complexity for group size 1, of the SIC-LSE, varies linearly with K and is 33% that of the full JD for $K = 16$. Note that maximum number of bursts in the UTRA TDD system is 16. The complexity advantage of the SIC-JD over full JD will be even more pronounced when the exact Cholesky decomposition is employed. Exact Cholesky decomposition's complexity exhibits a stronger dependence on K, leading to more savings as the dimension of the JD is reduced via SIC-JD.

Total number of bursts	Complexity of the SIC-JD expressed as a percentage of the complexity of the single-step JD of all K bursts			
	K groups of size 1 each (SIC-LSE)	K / 2 groups of size 2 each	K / 4 groups of size 4 each	K / 8 groups of size 8 each
8	63 %	67 %	76 %	100 %
16	33 %	36 %	41 %	57 %

Table 1

As shown in Table 1, when the number and size of codes is made completely adaptive on an observation interval-by-observation interval basis, the SIC-JD provides savings, on average, over full JD. Since, on average, all bursts do not arrive at the receiver with equal power, depending upon the grouping threshold, the size of the groups will be less then the total number of arriving bursts. In addition, a reduction in peak complexity is also possible if the maximum allowed group size is hard-limited to be less than the maximum possible number of bursts. Such a scheme leads to some degradation in performance when the number of bursts arriving at the receiver with the roughly the same power exceeds the maximum allowed group size. Accordingly, SIC-JD provides a mechanism to trade-off performance with peak complexity or required peak processing power.

* * *

CLAIMS

What is claimed is:

1. A method for use in receiving a plurality of data signals transmitted over a shared spectrum in a time slot in a time division duplex communication system using code division multiple access, the method comprising:

receiving a combined signal over the shared spectrum in the time slot;

grouping the plurality of data signals into a plurality of groups;

match filtering the combined signal based on in part symbol responses associated with the data signals of one of the groups;

jointly detecting data from each data signal in the one group;

constructing an interference signal based on in part the one group detected data;

subtracting the constructed interference signal from the combined signal; and

determining data from a group other than the one group by processing the subtracted signal.

2. The method of claim 1 wherein the jointly detecting is performed using least squares estimation.

3. The method of claim 1 wherein the jointly detecting is performed using minimum mean square error estimation.

4. A method for use in receiving a plurality of data signals transmitted over a shared spectrum in a time slot in a time division duplex communication system using code division multiple access, the method comprising:

(a) receiving a combined signal as an input signal over the shared spectrum in the time slot;

(b) grouping the plurality of data signals into a plurality of groups, at least one of the groups having a plurality of data signals;

(c) match filtering the input signal based on in part symbol responses associated with each data signal of a first group of the groups;

(d) jointly detecting data from each data signal in the first group;

(e) constructing an interference signal based on in part the first group detected data;

(f) subtracting the constructed interference signal from the input signal as an input signal for subsequent processing;

(g) match filtering the subtracted signal based on in part symbol responses associated with the data signal of a subsequent group of the groups;

(h) jointly detecting data from each data signal in the subsequent group; and

(i) successively repeating steps (e) through (h) for remaining groups of the plurality of groups where, for each remaining group, the subsequent group acts as the first group for that remaining group and that remaining group acts as the subsequent group.

5. A method for use in a receiver for receiving a plurality of data signals transmitted over a shared spectrum in a time slot in a time division duplex communication system using code division multiple access, the method comprising:

receiving a combined signal over the shared spectrum in the time slot;

estimating a received power level for each data signal;

selectively grouping data signals of the plurality of data signals based on in part the received power level of the data signals into at least one group; and

separately detecting data within each group for that group's data signals.

6. The method of claim 5 wherein the estimating the received power level for each data signal is based on in part apriori knowledge at the receiver.

7. The method of claim 5 wherein the estimating the received power level for each data signal is based on in part a power level of a training sequence associated with each data signal.

8. The method of claim 5 wherein the estimating the received power level for each data signal is performed using a bank of matched filters, each matched filter matched to a code of a respective one of the data signals.

9. The method of claim 5 wherein the selectively grouping data signals groups data signals within a certain threshold power level into a group.

10. The method of claim 9 wherein the certain threshold power level is one decibel.

11. The method of claim 9 wherein the certain threshold is adjusted to achieve a desired bit error rate at the receiver.

12. The method of claim 5 further comprising forcing all of the data signals into a single group to override the step of selectively grouping.

13. The method of claim 5 further comprising forcibly grouping each data signal into its own group to override the step of selectively grouping.

14. A method for use in a receiver for adjusting a trade-off between complexity and performance in detecting data from data signals transmitted over a shared spectrum in a time slot in a time division duplex communication system using code division multiple access, the method comprising:

grouping the data signals into at least one group; wherein to reduce the complexity, increasing a number of data signal groups, and to increase the performance, decreasing a number of data signal groups; and

jointly detecting data in each group.

15. The method of claim 14 further comprising:

determining a received power of each data signal; wherein the grouping is performed so that all data signals within each group are within a certain threshold power level and to reduce complexity, the certain threshold is increased and to increase performance, the certain threshold is reduced.

16. The method of claim 14 wherein to reduce the complexity, each group contains one of the data signals.

17. The method of claim 14 wherein to increase the performance, the at least one group is a single group.

18. A receiver for use in a time division duplex communication system using code division multiple access, the system communicating using multiple communication bursts in a time slot, the receiver comprising:

an antenna for receiving radio frequency signals including the multiple communication bursts;

a demodulator for demodulating radio frequency signals to produce a baseband signal;

a channel estimation device for estimating a channel response for the bursts;

a successive interference cancellation joint detection (SIC-JD) device comprising:

a first joint detection block for detecting data within the baseband signal for a first group of bursts of the multiple bursts;

a first interference construction block for constructing an estimate of interference of the first group bursts;

a subtractor for subtracting the first group interference from the baseband signal; and

a second joint detection block for detecting data within the subtracted signal for a second group of bursts of the multiple bursts.

19. The receiver of claim 18 wherein the SIC-JD device further comprises:

a plurality of additional joint detection blocks for detecting data for additional groups of bursts of the multiple bursts.

20. The receiver of claim 18 wherein the SIC-JD device further comprises:

a first matched filter for processing the baseband signal to match symbol responses of the data signals in the first group; and

a second matched filter for processing the subtracted signal to match symbol responses of the data signals in the second group.

21. The receiver of claim 18 wherein an output of the first and second joint detection blocks are soft symbols, the SIC-JD device further comprising a first and second soft to hard decision block for converting the first and second joint detection block outputs into hard symbols.

22. A device for use in a receiver of a time division duplex communication system using code division multiple access, the system communicating using multiple communication bursts in a time slot, the device comprising:

an input configured to receive a baseband signal associated with received bursts within a time slot;

a first joint detection block for detecting data within the baseband signal for a first group of bursts of the received bursts;

a first interference construction block for constructing an estimate of interference of the first group bursts;

a subtractor for subtracting the first group interference from the baseband signal; and

a second joint detection block for detecting data within the subtracted signal for a second group of bursts of the received bursts.

23. The device of claim 22 further comprising additional joint detection blocks for detecting data for additional groups of bursts of the multiple bursts.

24. The device of claim 22 further comprising:

a first matched filter for processing the baseband signal to match symbol responses of the received bursts of the first group; and

a second matched filter for processing the subtracted signal to match symbol responses of the received bursts of the second group.

25. The device of claim 22 wherein an output of the first and second joint detection blocks are soft symbols, the device further comprising a first and second soft to hard decision block converting the first and second joint detection block outputs into hard symbols.

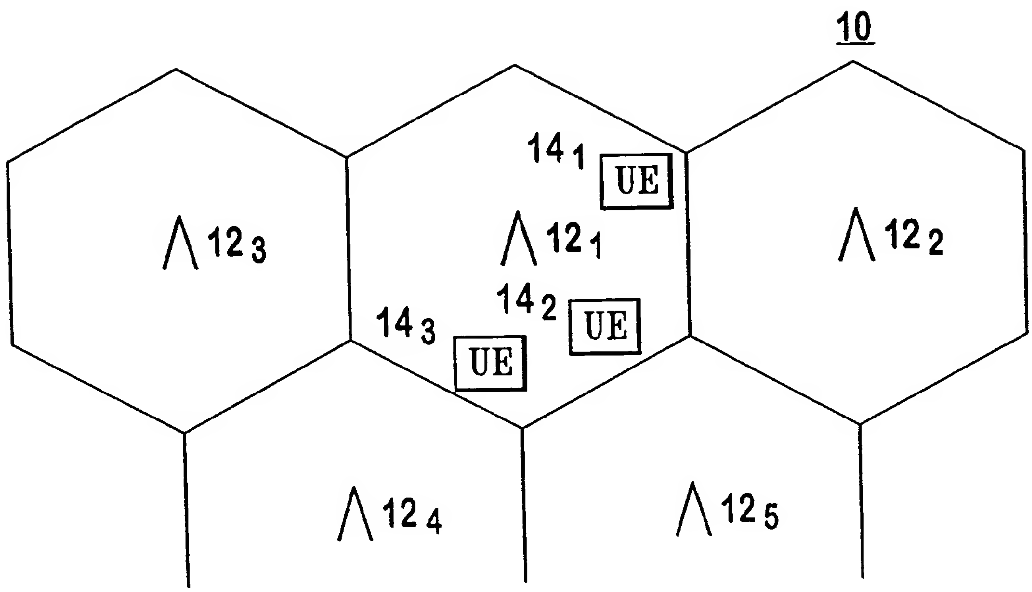

10
14 1
UE
12 3
12 1
12 2
14 2
14 3
UE
UE
12 4
12 5

FIG. 1

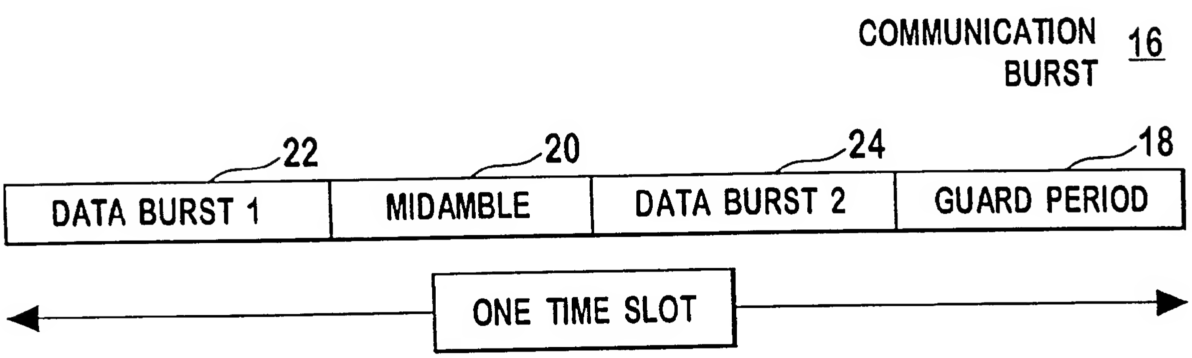

COMMUNICATION BURST
16
22
20
24
18
DATA BURST 1
MIDAMBLE
DATA BURST 2
GUARD PERIOD
ONE TIME SLOT

FIG. 3

2/8

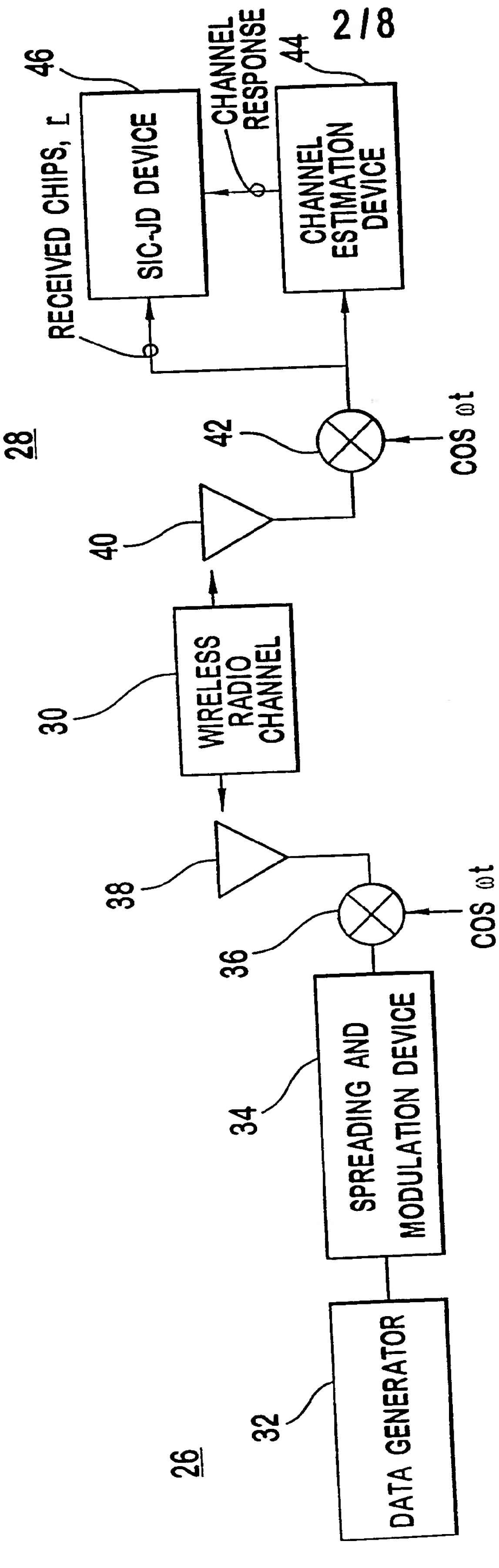

26
DATA GENERATOR
32
SPREADING AND MODULATION DEVICE
34
36
COS ωt
38
30
WIRELESS RADIO CHANNEL
40
42
COS ωt
28
RECEIVED CHIPS, r
SIC-JD DEVICE
46
CHANNEL RESPONSE
CHANNEL ESTIMATION DEVICE
44

FIG. 2

48 GROUP THE BURSTS, 1 TO K, INTO G GROUPS

50 MATCH FILTER THE RECEIVED SIGNAL, $\underline{x}_g^{(1)} = r$, with THE SYMBOL RESPONSES OF GROUP 1, PRODUCING $\underline{y}_g^{(1)}$

52 JOINTLY DETECT SOFT SYMBOLS WITHIN THE MATCH FILTERED SIGNAL, $\underline{y}_g^{(1)}$, FOR THE GROUP 1 BURSTS

54 CONVERT THE SOFT SYMBOLS INTO HARD SYMBOLS FOR THE GROUP 1 BURSTS

56 CONSTRUCT AN INTERFERENCE SIGNAL USING THE GROUP 1 DETECTED HARD SYMBOLS

58 SUBTRACT THE GROUP 1 INTERFERENCE SIGNAL FROM THE RECEIVED SIGNAL TO RESULT IN A GROUP 1 INTERFERENCE CANCELLED SIGNAL FOR GROUP 2, $\underline{x}_g^{(1)} - \hat{\underline{r}}^{(1)} = \underline{x}_g^{(2)}$

60 PROCESS GROUP 2 SIMILARLY TO GROUP 1

62 SUBTRACT THE GROUP 2 INTERFERENCE SIGNAL, $\hat{\underline{r}}^{(2)}$, FROM THE CANCELLED SIGNAL FOR GROUP 2, $\underline{x}_g^{(2)} - \hat{\underline{r}}^{(2)} = \underline{x}_g^{(3)}$

64 PROCESS THE REMAINING GROUPS, 3 ··· G, SIMILARLY, EXCEPT INTERFERENCE CONSTRUCTION IS NOT PERFORMED USING THE GROUP G HARD SYMBOLS

FIG. 4

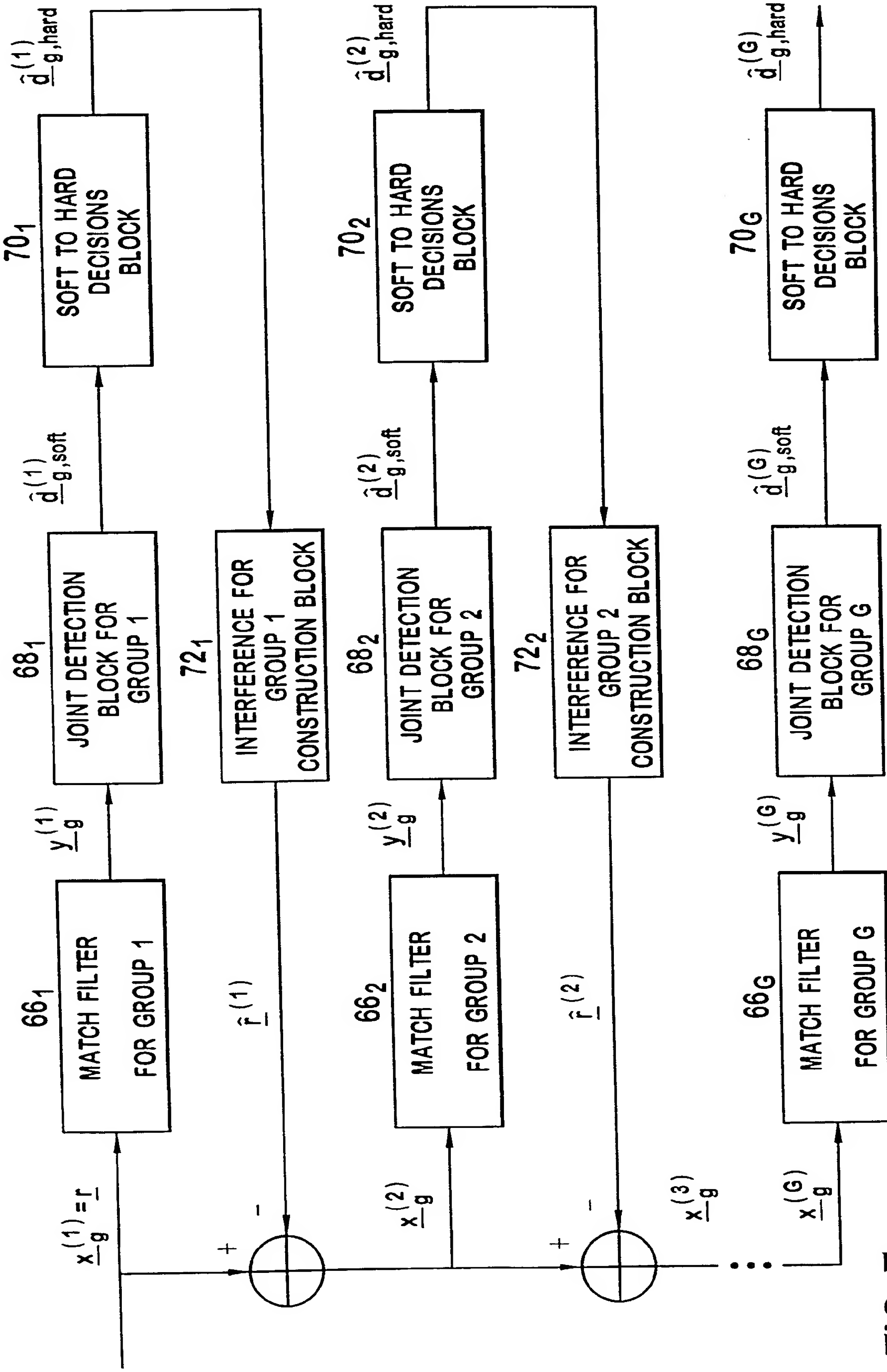

$\underline{x}_g^{(1)}=\underline{r}$
66_1
MATCH FILTER FOR GROUP 1
$\underline{y}_g^{(1)}$
68_1
JOINT DETECTION BLOCK FOR GROUP 1
$\hat{\underline{d}}_{g,soft}^{(1)}$
70_1
SOFT TO HARD DECISIONS BLOCK
$\hat{\underline{d}}_{g,hard}^{(1)}$
72_1
INTERFERENCE FOR GROUP 1 CONSTRUCTION BLOCK
$\hat{\underline{r}}^{(1)}$
+
-
$\underline{x}_g^{(2)}$
66_2
MATCH FILTER FOR GROUP 2
$\underline{y}_g^{(2)}$
68_2
JOINT DETECTION BLOCK FOR GROUP 2
$\hat{\underline{d}}_{g,soft}^{(2)}$
70_2
SOFT TO HARD DECISIONS BLOCK
$\hat{\underline{d}}_{g,hard}^{(2)}$
72_2
INTERFERENCE FOR GROUP 2 CONSTRUCTION BLOCK
$\hat{\underline{r}}^{(2)}$
+
-
$\underline{x}_g^{(3)}$
$\underline{x}_g^{(G)}$
66_G
MATCH FILTER FOR GROUP G
$\underline{y}_g^{(G)}$
68_G
JOINT DETECTION BLOCK FOR GROUP G
$\hat{\underline{d}}_{g,soft}^{(G)}$
70_G
SOFT TO HARD DECISIONS BLOCK
$\hat{\underline{d}}_{g,hard}^{(G)}$

FIG. 5

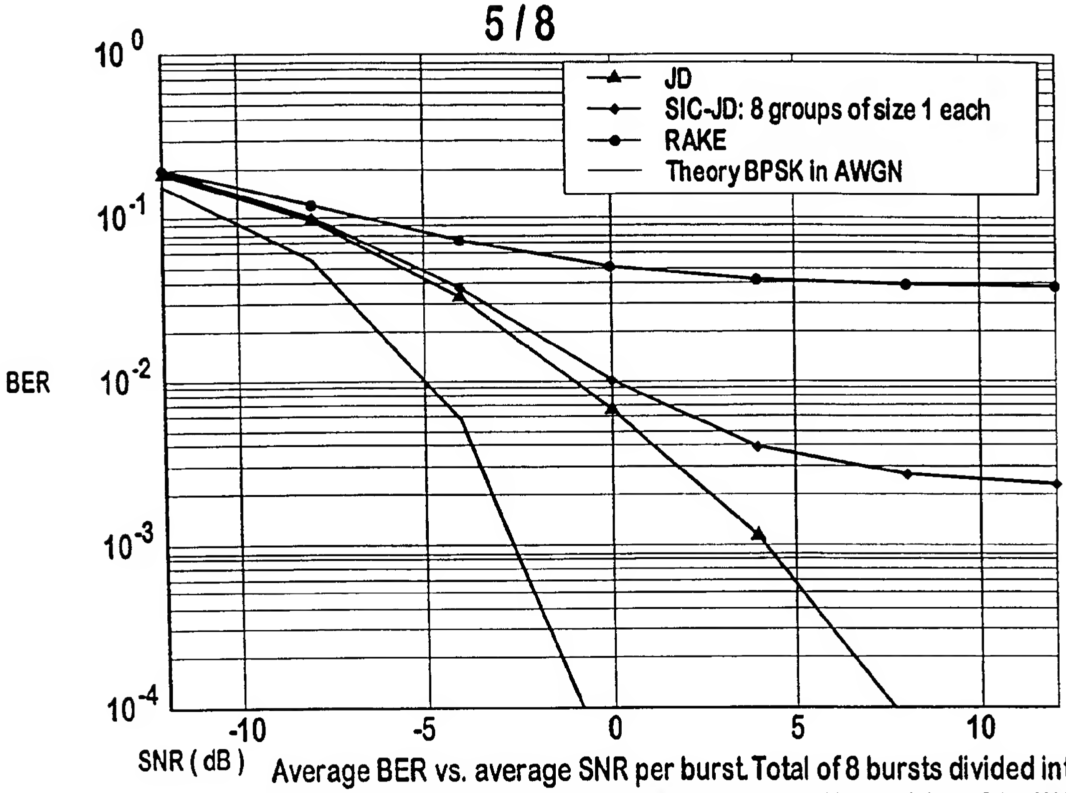


FIG. 6

Average BER vs. average SNR per burst. Total of 8 bursts divided into 8 groups with 1 burst per group. Multi-path profile is of the 3GPP WG4 Case 2 type. All 8 bursts have the same average SNR but pass through uncorrelated channel.

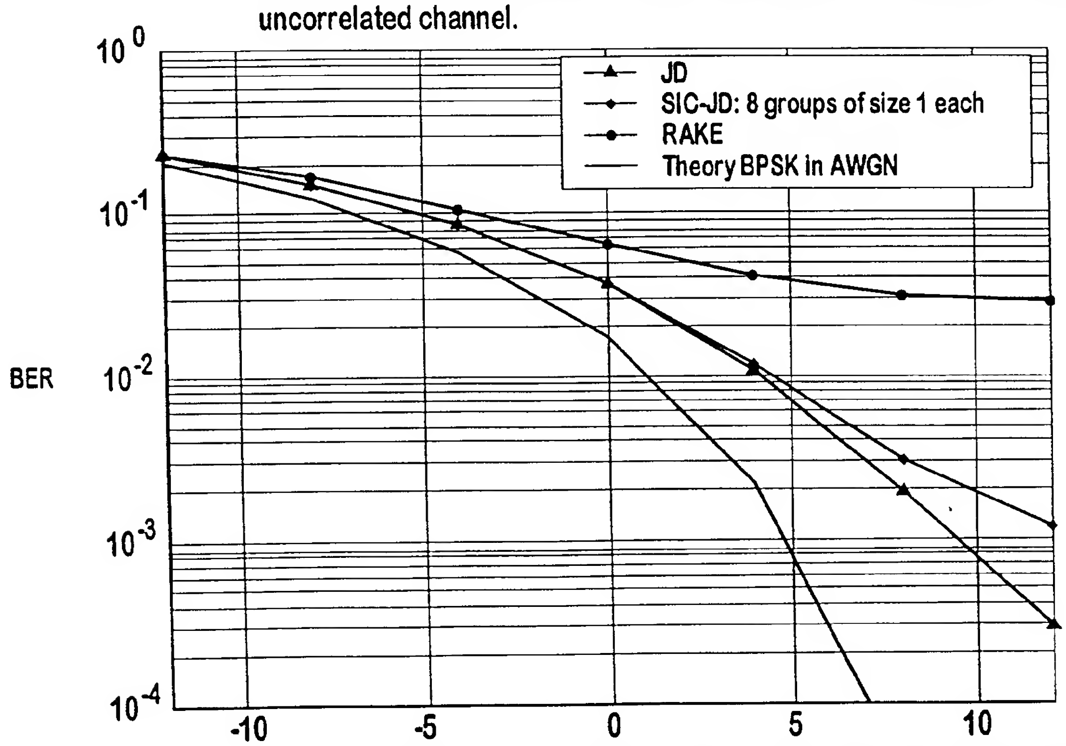
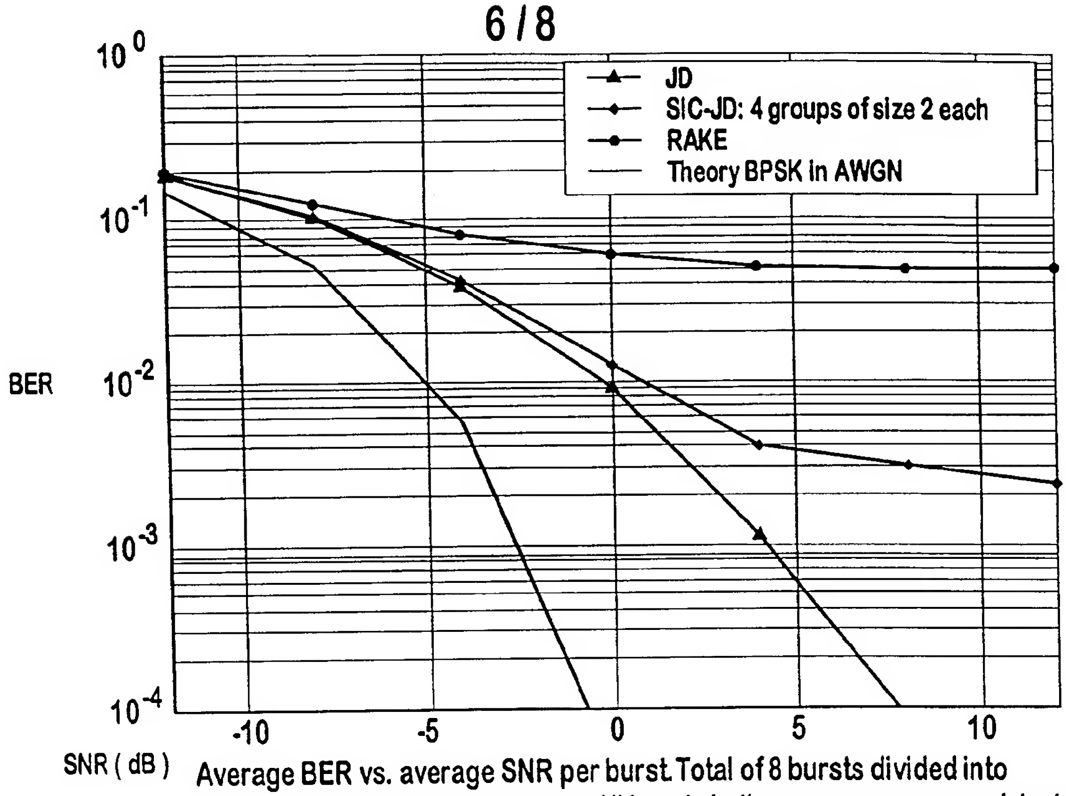


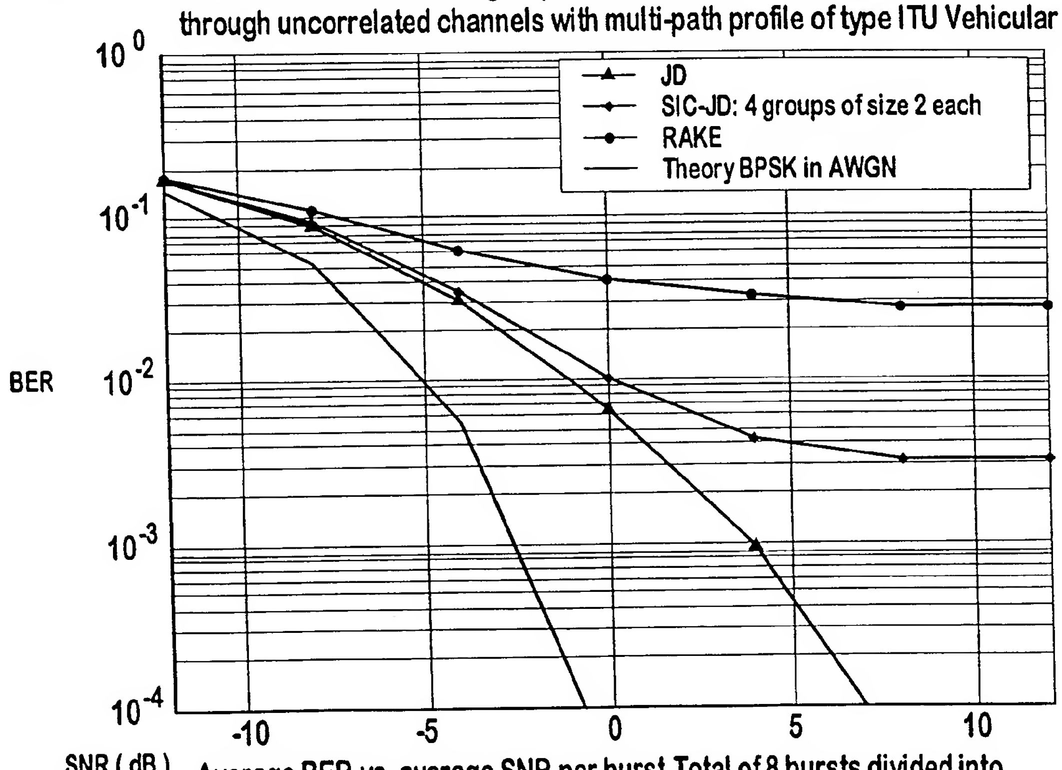
FIG. 7

Average BER vs. average SNR per burst. Total of 8 bursts divided into 8 groups with 1 burst per group. Multi-path profile is of the 3GPP WG4 Case 2 type. All 8 bursts pass through a common channel but their average SNR is separated by 2 dB.



Average BER vs. average SNR per burst. Total of 8 bursts divided into 4 groups with 2 bursts per group. All bursts in the same group are subjected to the same channel. All 4 groups have the same average SNR but pass through uncorrelated channels with multi-path profile of type ITU Vehicular A.

FIG. 8



Average BER vs. average SNR per burst. Total of 8 bursts divided into 4 groups with 2 bursts per group. All bursts in the same group are subjected to the same channel. All 4 groups have the same average SNR but pass through uncorrelated channels with multi-path profile of type 3GPP WG4 Case 2.

FIG. 9

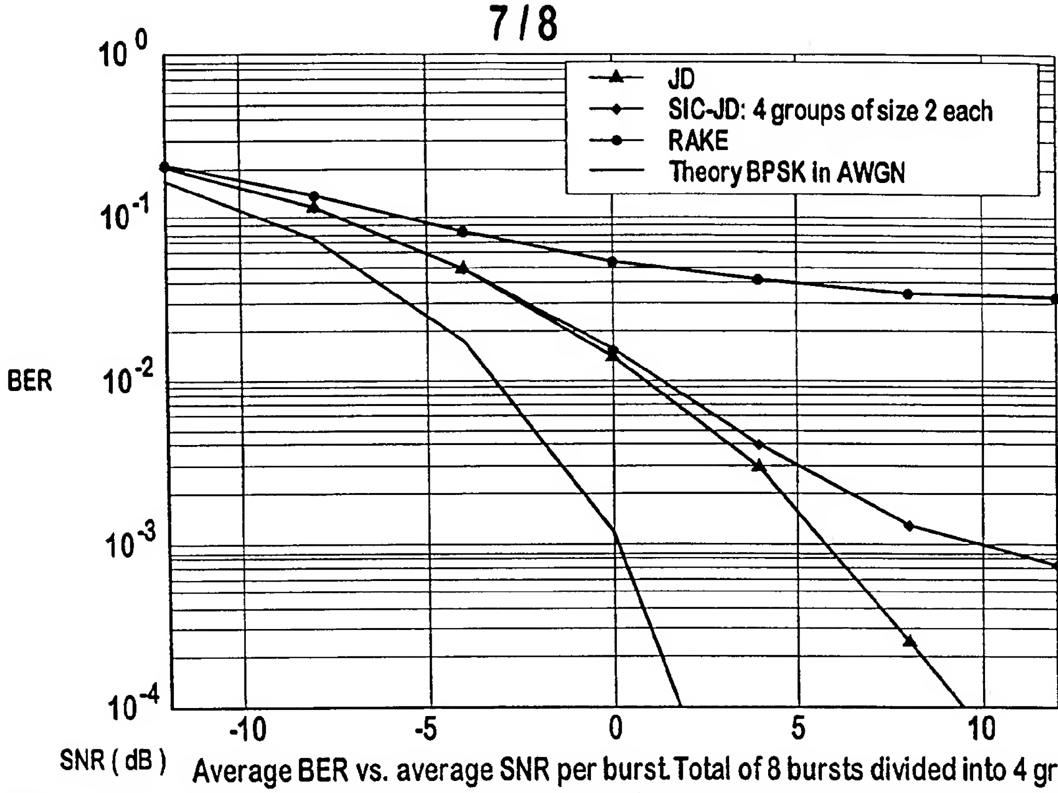


FIG. 10 Average BER vs. average SNR per burst. Total of 8 bursts divided into 4 groups with 2 bursts per group. All bursts in the same group are subjected to the same channel. All 4 groups pass through a common channel but their average SNR is seperated by 2 dB. Multi-path profile is of type ITU Vehicular A.

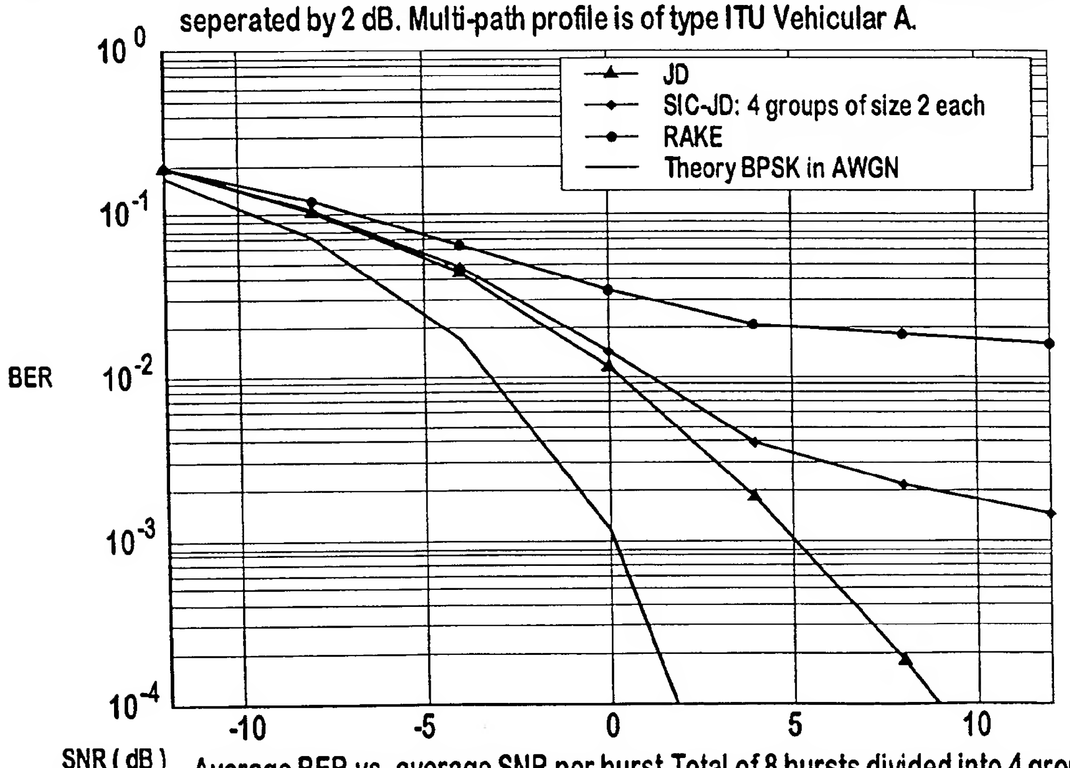


FIG. 11 Average BER vs. average SNR per burst. Total of 8 bursts divided into 4 groups with 2 bursts per group. All bursts in the same group are subjected to the same channel. All 4 groups pass through a common channel but their average SNR is seperated by 2 dB. Multi-path profile is of type 3GPP WG4 Case 2.

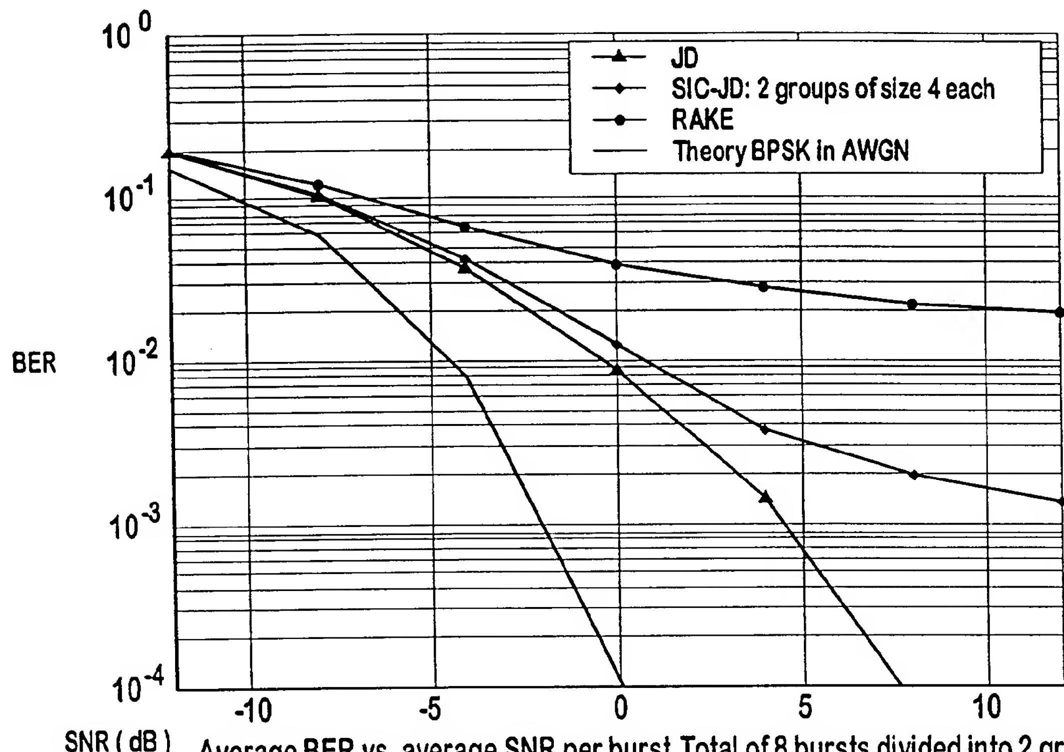


FIG. 12 Average BER vs. average SNR per burst. Total of 8 bursts divided into 2 groups with 4 bursts per group. All bursts in the same group are subjected to the same channel. All groups pass through a common channel but their average SNR is seperated by 2 dB. Multi-path profile is of type ITU Vehicular A.

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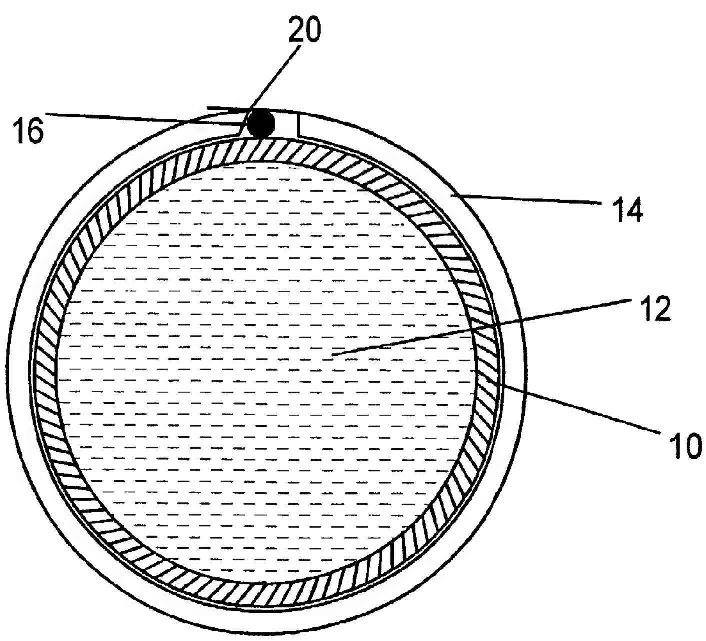
(54) Title: METHOD AND APPARATUS FOR LEAK DETECTION AND LOCATION



(57) Abstract: Leaks are detected by wrapping a vessel such as a pipe (10) or tank (28) in a skin (14, 14A, 30) which traps escaping fluid (12) at least long enough to direct it, in the vicinity of the leak, towards a sensor line (16) employing fibre optics to detect the fact of a leak and to detect how far along the a fibre optic line the leak exists. The skin (14, 14A, 30) can be longitudinally applied or wrapped onto the vessel (10, 28). Bindings (24) can be used to attach the skin (14, 14A, 30) to the outside of the vessel (10, 28). Sensor line (16) couplings (26) can be employed between lengths of pipe (10) to create monitored sections of pipe (10) which can be joined together. Sensor lines (16) can be applied to the outer surface of a vessel (10, 28) and covered with the skin (14, 14A, 30). Sensor lines (16) can be stuck or woven into the skin (14, 14A, 30) incorporates elastic ridges (18, 18A, 18C) which face the vessel and direct escaping fluid towards the sensor line (16). A control system (46, 40, 48, 44) is provided to shut down a tank (28) or pipeline (36, 38), at least in the vicinity of a leak, if a leak is detected, and can include shutting down pumps (40), closing valves, and voiding items (10, 28) subject to the leak.

WO 02/088656 A1

— *before the expiration of the time limit for amending the claims and to be republished in the event of receipt of amendments*

For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

Method and Apparatus for Leak Detection and Location

The present invention relates to a method and apparatus for detecting leaks. The invention particularly relates to detecting fluid (liquid or gaseous) leaks in vessels such as pipelines and storage tanks.

Pipelines are vessels used for conducting fluids, such as gas, water, chemicals or oil, from geographical place to geographical place, or, in an industrial setting, between tanks, between processes, or between processes and tanks and vice versa. Tanks are vessels used for temporary or permanent bulk storage, where the fluid enjoys at least a temporary period of non-movement. There is often a need to detect whether or not a pipe or tank is leaking. Often, this is done by visual inspection, by which time any damage done by the escaping fluid has already occurred, or is done by flow measurements where it is noted that the ingress of fluid volume or quantity is greater than the egress of fluid volume or quantity. In a tank, a leak becomes apparent either by inspection or by noting that the content of the tank has decreased over the sum of its inflow and outflow. The prior art is silent upon any method which will detect a leak, as it occurs anywhere in the system or tank under surveillance. The prior art is also silent upon any method automatically to locate the position of a leak as it occurs.

The present invention seeks to provide a method and apparatus for the rapid detection of the occurrence of a leak anywhere in a vessel or system of vessels under surveillance and the rapid determination of the location of the leak. The present invention also seeks to provide a method and apparatus whereby a leak, anywhere in a system of vessels under surveillance, can be shut down at the moment of detection.

According to a first aspect, the present invention consists in an apparatus for determining the occasion and location of a fluid leak in a vessel, said apparatus comprising: a sensor line, on the outside of the vessel, for detecting where, along the length of the sensor line the sensor line is in contact with the fluid; and a skin, for use on the outside of the vessel and operative to direct leaked fluid towards the sensor line in the vicinity of any leak.

According to a second aspect, the present invention consists in a method for determining the occasion and location of a fluid leak in a vessel, said method including the steps of: disposing a sensor line, on the outside of the vessel, for detecting where, along the length of the sensor line the sensor line is in contact with the fluid; disposing a skin on the outside of the vessel; and employing said skin to direct leaked fluid towards the sensor line in the vicinity of any leak.

The invention further provides that the said skin is operative, at least temporarily, to contain fluid, leaked from the vessel.

The invention further provides that the skin is operative to enclose both the vessel and the sensor line.

The invention further provides that the vessel can be a pipe, that the sensor line is disposable longitudinally along the pipe, that the skin comprises elastic ridges, to be pressed against the outer surface of the pipe, that the elastic ridges are operative to inhibit fluid flow in a longitudinal direction along the outside of the pipe, and that the elastic ridges are operative to direct fluid flow on the outside of the pipe in a circumferential direction towards the sensor line.

The invention further provides that the skin can be disposed longitudinally along the outside of the pipe and can be closed, longitudinally, by means of a securing cover

The invention further provides that the skin can be helically wrapped about the pipe.

The invention further provides that the skin can be further fixed onto the outside of the pipe by means of spaced bands.

The invention further provides that the vessel can be a tank, that the sensor line can be disposed on the outer surface of the tank, that the skin can comprise elastic ridges, to be pressed against the outer surface of the tank, that the elastic ridges can be operative to form containment zones to contain and accumulate fluid from any leak for the fluid to come into contact with the sensor line.

The invention further provides that the skin can be wrapped around the outside of the tank in a close helix, that the skin can comprise partial containment zones, and that a partial containment zone in one coil of the helix can co-operate with a partial containment zone in an adjacent coil of the helix to contain and accumulate fluid from any leak for the fluid to come into contact with the sensor line.

The invention further provides that the sensor line can be incorporated into the fabric of the face of the skin which is for presentation to the outside surface of the vessel.

The invention further provides that the sensor line can be a fibre optic line.

The invention is further explained, by way of example, by the following description, in conjunction with the appended

drawings, in which:

Figure 1 shows a cross sectional view of a vessel in the form of a pipe to which the present invention has been applied.

Figure 2 shows an opened out view of the skin enclosing the pipe of figure 1.

Figure 3 shows one way of attaching the skin, of Figures 1 & 2, to a pipe.

Figure 4 shows another way that a skin may be applied to a pipe.

Figure 5 is an example of the skin, which can be used in Figure 4.

Figure 6 is an angled view of the pipe, shown in Figure 1.

Figure 7 illustrates how sensor lines may be incorporated into the skin otherwise shown in Figure 2.

Figure 8 is an example of how sensor lines may be incorporated into the skin, otherwise shown in Figure 5.

Figure 9 is a drawing of a vessel in the form of a tank showing how a skin may be applied thereto, not only to detect a leak, but to determine at what part of the tank the leak is occurring.

Figure 10 is a view of the ridge structure of a skin suitable for use on the tank of Figure 9, and showing how sensor lines may be applied thereto.

Figure 11 is a cross sectional view of a pipeline showing how sensor lines can be attached other than at the top of a pipe and how the skin need not provide containment of leaking fluid, merely direction toward the sensor line.

And

Figure 12 is a projected schematic view of an exemplary pumping system, according to the present invention, showing a control system suitable for monitoring a pipeline for leaks, for shutting down the pipeline when a leak is detected, and for providing rapid indication of the location of the leak.

Attention is drawn to Figure 1. A pipe 10, carrying a fluid load 12, is surrounded by a containment skin 14. Within the containment skin 14, and on top of the pipe 10, a sensor line 16 is provided.

The fluid load 12 can be gaseous or liquid. It can consist of chemical gases, fuel gases, hydrocarbons, oil, water, food stuffs such as milk, chemical liquids and, indeed, just about any type of thing that can be driven along a pipe 10.

Part of the purpose of the skin 14 is to protect the sensor line 16 when the pipe 10 is buried in the ground, encased in concrete, or otherwise exposed to a harsh surrounding environment. Another purpose of the containment skin 14 is, at least temporarily, to contain any of the fluid load 12 which may escape from the pipe 10 at least long enough to duct the escaping fluid load 12 towards the sensor line 16.

The sensor line 16 is, in the preferred embodiment of this invention, a fibre optic line which, as is well known to those skilled in the art, can be adapted and used to detect, among other things, moisture, specific chemicals, changes in

temperature, oil and natural gas. The present invention can also employ any other elongated sensor or array of sensors, including spaced gas, chemical oil, temperature and other sensors. More than one sensor line 16 can be provided.

By placing the sensor line 16 on top of the pipe 10, the sensor line 16 avoids contact with accumulated contaminants and debris which might accrete in the bottom of the skin 14, and ensures, thereby, that the sensor line 16 responds only to true leaks. As will become clear from the description of Figure 11, other arrangements are possible within the scope of the present invention.

Attention is drawn to Figure 2, showing the face of the skin 14 applied to the pipe 10. The reverse face of the skin 14 is smooth. The skin 14 comprises elastic ridges 18 which are wrapped circumferentially around the pipe 10 and held in place by a securing cover 20 which can be used to close the skin 14 by means of adhesives or other gripping means, and also serves to protect the sensor line 16 and to maintain it in position on top of the pipe 10.

Should a leak occur, the circumferential elastic ridges 18, against the face of the pipe 10, prevent the escaping fluid 12 from moving longitudinally along the pipe 10. Any fluid escape is ducted substantially circumferentially around the pipe at the location where it occurred. As is known in the art, a fibre optic sensor line 16 can detect the position along its length where interaction with a selected or detectable fluid has occurred. By confining the leaking fluid 12 to the point where the leak occurred, at least long enough for the escaping fluid to encounter the sensor line 16, and by sensing the position of interaction of the fibre optic sensor line 16 with the fluid, ducted towards the fibre optic sensor line 16, it is possible to obtain a very rapid detection of

the fact that a leak has occurred and to find the position of that leak, with great accuracy. Distance along the fibre optic sensor line 16 is measured by finding the time delay for light travelling along the fibre optic sensor line 16.

The lower portion of Figure 2 is a side view of the skin 14, looking in the direction of the arrow 22, showing the elastic ridges 18 in profile.

Attention is drawn to Figure 3 showing one way in which the skin 14 of Figure 1 can be attached to a pipe 10 (shown in phantom outline) by means of spaced bands 24 braced over the skin 14 at intervals along the pipe 10, not only to hold the skin 14 onto the pipe 10 but also to improve the ability of the skin 10 longitudinally to trap escaping fluid from the pipe 10. Couplings 26 allow the sensor line 16 to be coupled to sensor lines 16 on adjacent pipes 10. The arrangement shown in Figure 3 constitutes a complete and portable pipe 10 unit which can be moved and installed as an entirety.

Attention is drawn to Figure 4 showing another way in which a spiral skin 14A can be helically wrapped around the pipe 10 (shown in phantom outline).
Figure 5 shows the arrangement of elastic ridges 18A on the spiral skin 14A. It is perceived that the elastic ridges 18A are longitudinal in the sense of direction of the spiral skin 14A, and when wrapped around the pipe 10 in a close fitting helical fashion, the elastic ridges 18A in the spiral skin 14A form a more or less vertical (circumferential) pattern which contains any leak in the vicinity of the sensor line 16. In the example shown in Figure 4, the sensor line 16 is simply placed on top of the pipe 10 and the spiral skin 14A wrapped around the pipe 10. As will later be seen, better arrangements than this can be made. The elastic ridges 18, 18A can be provided at various angles to the longitudinal

direction of the skin 14, dependently upon the intended manner of attachment thereof to the pipe 10.

The arrangement shown in Figure 4 can further be improved by, in addition to helical wrapping, employing spaced bands 24 as illustrated in Figure 3.

Attention is drawn to Figure 6, showing, for clarity, a projected view of the cross section of Figure 1, and showing, in particular, how the sensor line 16 is enclosed by and protected by the securing cover 20. The containment skin 14, shown in Figure 6, is shown cut away to cover only a portion of the surface of the pipe 10 so that the disposition of the sensor line 16 can be seen. While the sensor line 16 is shown covered by the securing cover 20, it is to be understood that the skin 14 can be secured to the pipe 10 by other means and the sensor line 16 left uncovered by the skin 14 simply to have any escaping fluid 12 ducted in its direction.

Attention is drawn to Figure 7 showing one way in which one or more sensor lines 16 can be threaded, permanently, through the material or fabric on the inside of the containment skin 14, otherwise shown in Figure 2. If the inner face of the containment skin 14 contains any woven fabric element, the sensor line 16 can simply be woven into the fabric. Otherwise, the sensor line 16 can simply be moulded into or attached to the inner surface of the containment skin 14.

Figure 8 shows a manner in which sensor lines 16 can be applied as an integral part of the spiral skin 14A shown in Figures 4 and 5. One or more sensor lines 16 are placed on the inner surface of the spiral skin 14A to run in the spaces between the elastic ridges 18A in the spiral skin 14A. They can be incorporated in just the same manner as was earlier described for Figure 7.

Attention is next drawn to Figure 9 showing an application of the present invention to a tank 28 containing a fluid. A tank skin 30 is wrapped around the tank 28, here shown wrapped in a spiral manner but other manners are possible, to cover the surface of the tank 28 as much as possible. The sensor line 16 leads all around the helical wrapping of the tank skin 30 and is accessible at either end via couplings 26A.

Attention is drawn to Figure 10 showing the face of the tank skin 30 which faces the tank 28. Elastic ridges 18C form entire containment zones 32 in the centre of the tank skin 30 and partial containment zones 34 at the sides thereof and extending on either side, away from the entire containment zone 32 to the edge of the tank skin 30. When pressed against the tank 28, the entire containment zones 32 keep any escaping fluid, at least temporarily, from moving. The partial containment zones 34 co-operate with partial containment zones in adjacent wraps of the tank skin 30 to form at least a temporary containment area for escaping fluid. The sensor line 16, in the example shown, is provided only through the entire containment zone 32 and through one of the partial containment zones 34. As the tank skin 30 is wrapped, the sensor line 16 in the adjacent wrap of tank skin 30 acts to provide a sensor line 16 in a co-operation of partial containment zones where one is not present on the adjacent tank wrap 30. Since it is possible to measure the distance along the fibre optic sensor line 16, it is possible to measure where, on the surface of the tank 28 a leak has occurred. By knowing which portion of the tank 28 is covered by which portion of the tank skin 30, the location of a leak can be rapidly determined.

Attention is next drawn to Figure 11, showing the cross sectional pipe arrangement of Figure 1, but with a different arrangement for the relative position of the sensor line 16.

The sensor line 16 is, for preference, provided at the top of the pipe 10. However, as is clear if more than one sensor line is used, the sensor line 16 may be otherwise disposed on the pipe 10. Figure 11 shows the sensor line 16 in a first position near the base of the pipe 10. The sensor line 16 is also shown in a second position where it is part way up the pipe 10. The sensor line can be attached, in varying positions on the pipe 10 by means of adhesives, tapes and spaced bands. All that is important is that the sensor line is in a position to interact with the escaping fluid 12, should an escape occur. The sensor line 16 can be longitudinally disposed along the pipe, or can be spirally wound around the pipe 10 in an arrangement incorporating the arrangement of the spiral skin 14A of Figure 4 with an incorporated sensor lines 16 or lines 16 as shown in Figure 7 and Figure 8.

The elastic ridges 18 18A, shown in Figures 2, 5, 7, 8 and 10, can also be otherwise provided, according the present invention. The ridges 18 18A 18C may restrict the escaping fluid 12 from longitudinal migration along the pipe 10 . However, localisation of the escaping fluid 12 is only necessary for enough time for the fluid 12 to reach the sensor line16 . Thus, arrangements of ridges 18 18A 18C, right down to there being no ridges 18, 18A 18B present, merely a conformal skin that spreads the escaping fluid across the surface of the pipe 10 at least as far as the sensor line, can also be provided and can work according the the present invention.

The skin 14 need not contain escaping fluid. The invention also functions if the fluid 12 can escape. The skin, in Figure 11, shows a means of egress for the fluid, just to emphasize this point, where the securing cover 20 does not provide a fluid tight enclosure around the pipe 10.

Attention is drawn to Figure 12 showing a projected schematic diagram illustrating how the present invention provides a control feature for tanks or pipelines.

An exemplary pipeline 36 comprises one or more pipeline sections 38 between pumping stations 40 through which a fluid is propelled as illustrated by arrows 42. The pipeline 36 could equally well be a tank.

A sensor line 16D is provided for leak detection on one or more adjacent pipeline or tank sections 38 and is driven and monitored by a sensor driver 44 which provides laser light, laser detectors, timers and all the other apparatus which, as is already known in the art, is necessary for the detection and location of a fluid leak. A monitor 46 receives output from the sensor driver 44 and displays the current state of the monitored pipeline 36 or tank . As soon as the monitor 46 detects that a leak has occurred, it sends an operating signal to a pump controller 48 which sends a control signal to each pumping station 40 on the monitored pipeline 38 or tank causing each pumping station 40 to shut down. The monitor 46 can then provide humanly interpretable input for assessing the progress of leak repair and recovery.

The pump controller 48 could equally receive its operating signal directly from the sensor driver. The operating signal for the pumping stations 40 can be provided to all pumping stations 40 on the monitored and controlled pipeline 36or tank, or can be provided only to that pumping station 40 or those pumping stations 40 which is or are nearest to and contain the loss of fluid from the leak. In this example, the pipeline 36 or tank is provided with pumping stations. The invention provides that a pipeline or tank can comprise pumping stations, stop valves, and, indeed, any device which can be applied or ceased to be used in order to shut down the

loss of fluid flow from the pipeline 36 or tank whenever a leak is detected. This can comprise shutting down all flow. It can also comprises starting fluid movement out of the damaged section to prevent leakage loss, or a combination of both techniques. Thus, one pumping station 40 can be shut down in a pipeline 36 upstream of a leak and flow towards the leak stopped, while pumping downstream of the leak can be continued or enhanced to empty the pipeline 36. A monitored tank can have its inflow stopped while its outflow continues until the tank is empty, or until leakage stops as, for example, when the level in the lank falls below the height of the leak of until leakage is no longer detected.

The invention also provides that the sensor driver 44 can drive and monitor more than one sensor line 16D, either in the same pipeline (or tank) section 38 (where more than one indication of a leak can be employed to confirm a leak and prevent falsely indicated shutdowns) or in different pipeline (or tank) sections 38.

While it is implicit in the disclosure of the invention, it is here stated, for clarity, that the skin or skins 14 14A 30 can be flexible for wrapping around pipes, tanks and any other vessels to which the invention can be applied and that the invention, as described and claimed, can be retrofitted to existing pipes, tanks and other vessels.

The invention has so far been described by way of example. The invention is further described by the following claims.

Claims

1. An apparatus for determining the occasion and location of a fluid leak in a vessel, said apparatus comprising: a sensor line, on the outside of the vessel, for detecting where, along the length of the sensor line the sensor line is in contact with the fluid; and a skin, for use on the outside of the vessel and operative to direct leaked fluid towards the sensor line in the vicinity of any leak.

2. An apparatus, according to claim, wherein said skin is operative, at least temporarily, to contain fluid, leaked from the vessel.

3. An apparatus, according to claim 2, wherein said skin allows the escape of said fluid.

4. An apparatus, according to claims 1, 2 or 3, wherein said skin is operative to enclose both said vessel and said sensor line.

5. An apparatus, according to claims 1, 2, 3 or 4 for use where said vessel is a pipe, wherein said sensor line is disposable along the pipe

6. An apparatus, according to claim 5, wherein said skin comprises elastic ridges, to be pressed against the outer surface of the pipe and wherein said elastic ridges are operative to inhibit fluid flow in a longitudinal direction along the outside of the pipe and to directing fluid flow on the outside of the pipe towards said sensor line.

7. An apparatus, according to claim 5 or 6, wherein said skin is disposable longitudinally along the outside of the

pipe and closable, longitudinally, by means of a securing cover

8. An apparatus, according to claim 5 or 6, wherein said skin is helically wrappable about the pipe.

9. An apparatus according to claims 5, 6, 7 or 8, wherein said skin is further fixable onto the outside of the pipe by means of spaced bands.

10. An apparatus according to claims 1, 2, 3 or 4 for use where said vessel is a tank, wherein said sensor line is disposable on the outer surface of the tank.

11. An apparatus, according to claim 10, wherein said skin comprises elastic ridges, to be pressed against the outer surface of the tank, and wherein said elastic ridges are operative to form containment zones to at least temporarily contain fluid from any leak for the fluid to come into contact with said sensor line.

12. An apparatus, according to claims 10 or 11, wherein said skin is wrappable around the outside of the tank in a closed helix, wherein said skin comprises partial containment zones, and wherein a partial containment zone in one coil of the helix is co-operative with a partial containment zone in an adjacent coil of the helix to at least temporarily contain fluid from any leak for the fluid to come into contact with said sensor line.

13. An apparatus, according to any of the preceding claims, wherein said sensor line is incorporated into the fabric of the face of said skin which is for presentation to the outside surface of the vessel.

14. An apparatus, according to any of the preceding claims, wherein said sensor line is a fibre optic line.

15. An apparatus, according to any one of the preceding claims, including monitoring means, responsive to said sensor line to shut down said vessel in the event of detection of a leak.

16. An apparatus, according to claim 15, wherein said monitoring means is operative to cease pumping fluid at least into that portion of the vessel subject to the leak.

17. An apparatus, according to claims 15 or 16, wherein said monitoring means is operative to shut off at least that portion of the vessel subject to the leak.

18. An apparatus, according to any one of claims 15 to 17, wherein said monitoring means is operative to empty at least that portion of the vessel subject to the leak.

19. A method for determining the occasion and location of a fluid leak in a vessel, said method including the steps of: disposing a sensor line, on the outside of the vessel, for detecting where, along the length of the sensor line the sensor line is in contact with the fluid; disposing a skin on the outside of the vessel; and employing said skin to direct leaked fluid towards the sensor line in the vicinity of any leak.

20. A method, according to claim 19, including the step of employing said skin, at least temporarily, to contain fluid, leaked from the vessel.

21. A method according to claim 19, including the step of allowing the escape of said fluid.

22. A method, according to claims 19, 20 or 21, including the step of employing said skin to enclose both said vessel and said sensor line.

23. A method, according to any one of claims 19 to 22, where said vessel is a pipe, including the step of disposing said sensor line along the pipe.

24. A method, according to claim 23, including the steps of: employing elastic ridges on said skin; and pressing said elastic ridges against the outer surface of the pipe to inhibit fluid flow in a longitudinal direction along the outside of the pipe and to directing fluid flow on the outside of the pipe towards said sensor line.

25. A method, according to claims 23 or 24, including the steps of: disposing said skin longitudinally along the outside of the pipe; and closing said skin, longitudinally, by means of a securing cover

26. A method, according to claims 23 or 24, including the step of helically wrapping said skin about the pipe.

27. A method, according to any one of claims 23, 24, 25 or 26, including the step of fixing said skin onto the outside of the pipe by means of spaced bands.

28. A method, according to claims 19, 20, 21 or 22, where said vessel is a tank, said method including the steps of: disposing said sensor line on the outer surface of the tank.

29. A method, according to claim 28, including the steps of: employing elastic ridges on said skin; and pressing said elastic ridges against the outer surface of the tank to form containment zones to contain and accumulate fluid from any leak for the fluid to come into contact with said sensor line locally to the leak.

30. A method, according to claims 28 or 29, including the steps of: wrapping said skin around the outside of the tank in a close helix; and employing partial containment zones in said skin, a partial containment zone in one coil of the helix being co-operative with a partial containment zone in an adjacent coil of the helix to at least temporarily contain fluid from any leak for the fluid to come into contact with said sensor line locally to the leak.

31. A method, according to any one of claims 19 to 30, including the step of including said sensor line in the fabric of the face of said skin which is for presentation to the outside surface of the vessel.

32. A method, according to any one of claims 19 to 31, including the step of employing, in said sensor line, a fibre optic line.

33. A method, according to any one of claims 19 to 32, including the steps of monitoring said sensor line; and shutting down said vessel in the event of detection of a leak.

34. A method, according to claim 30, wherein said step of shutting down said vessel includes the step of ceasing to pump fluid at least into that portion of the vessel subject to the leak.

35. A method, according to claims 33 or 34, wherein said step of shutting down said vessel includes the step of shutting off at least that portion of the vessel subject to the leak.

36. A method, according to any one of claims 27 to 29, wherein said step of shutting down said vessel includes the step of emptying at least that portion of the vessel subject to the leak.

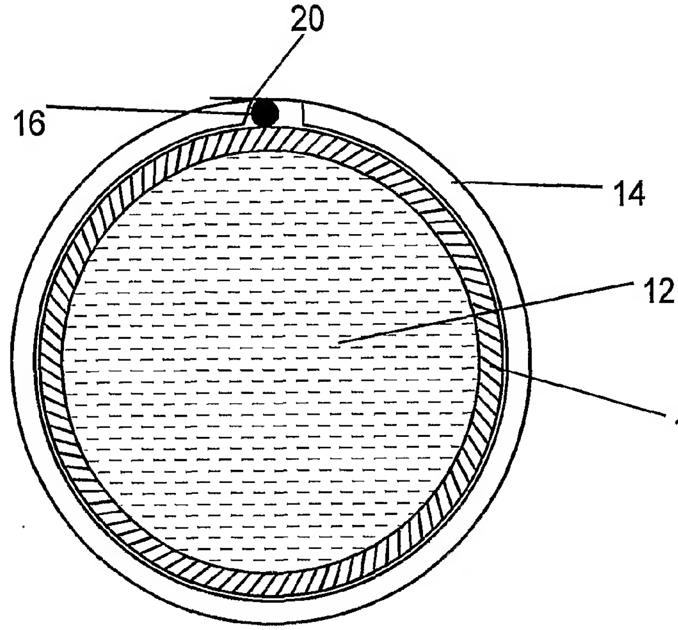


FIGURE 1

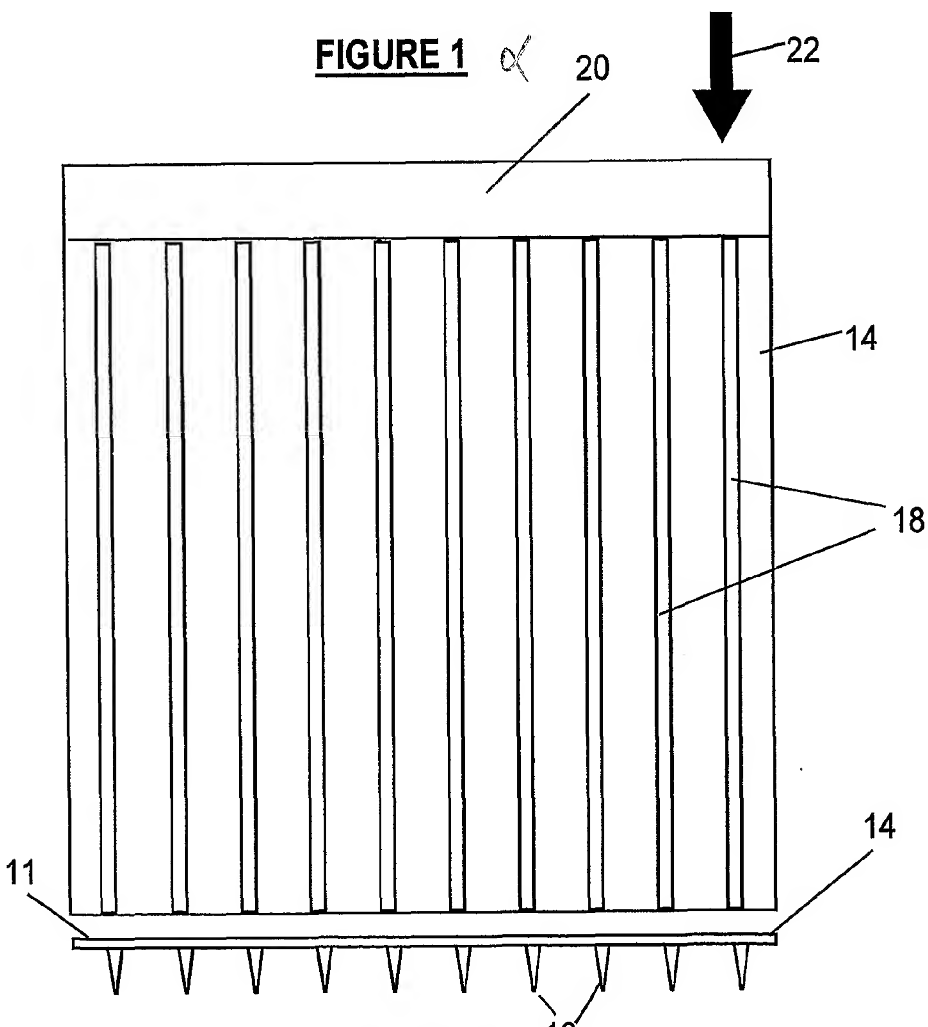


FIGURE 2

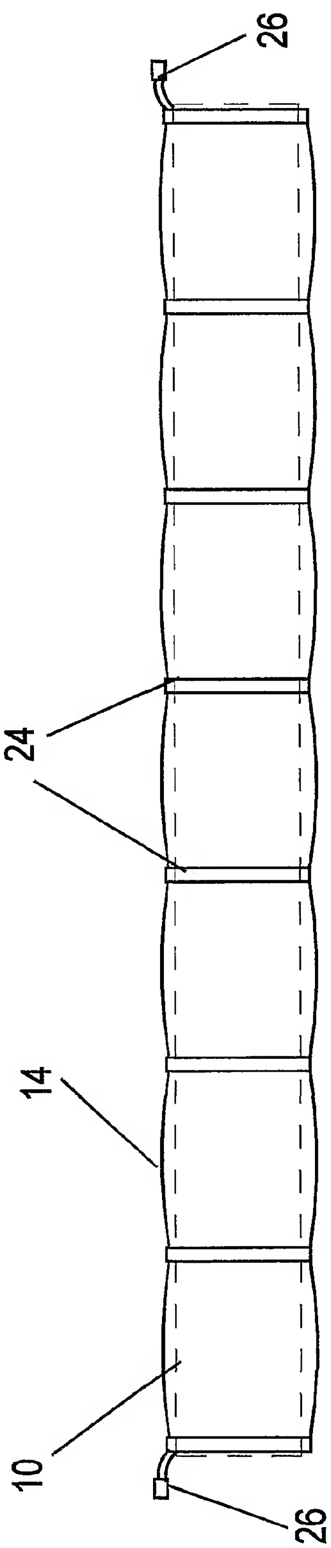


FIGURE 3

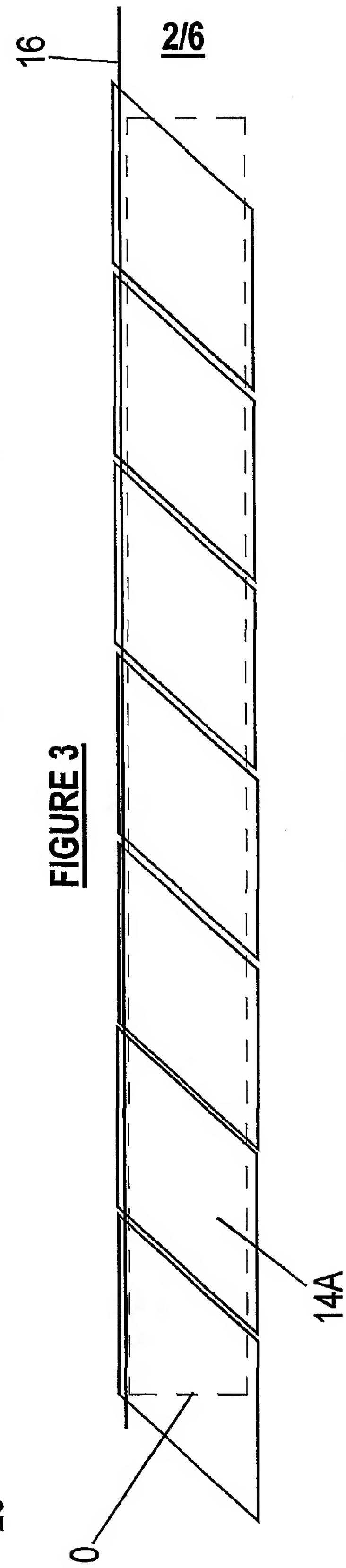


FIGURE 4

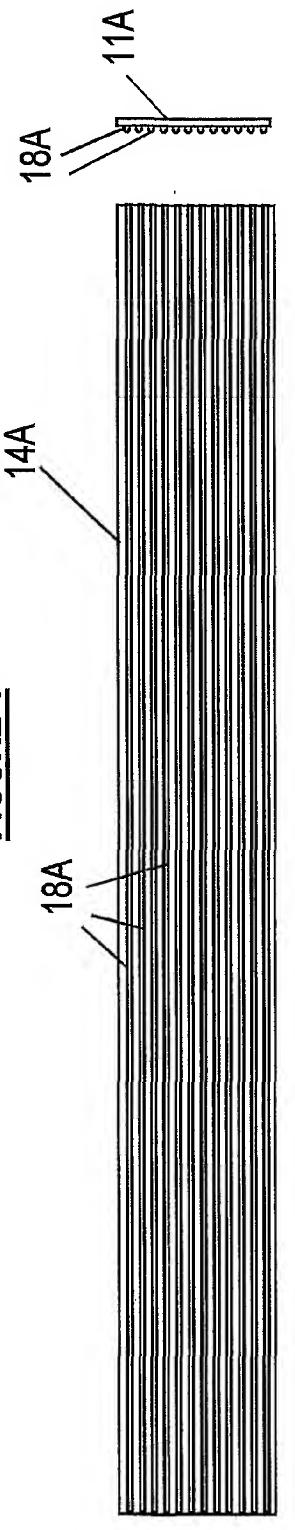


FIGURE 5

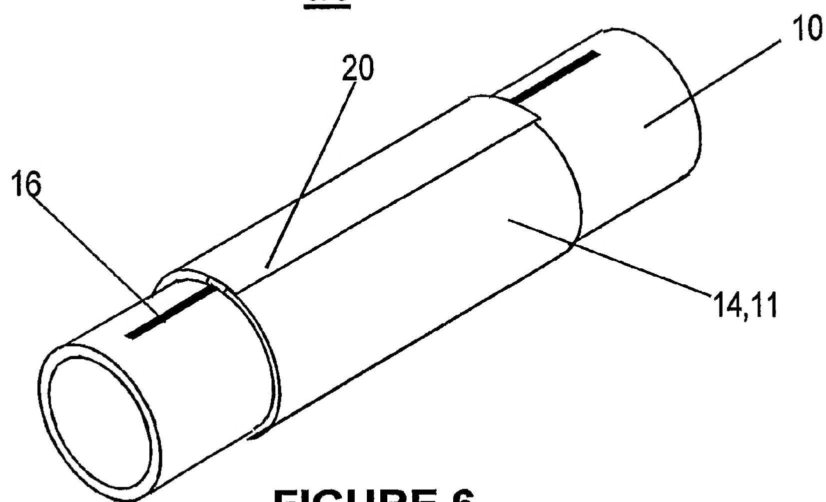


FIGURE 6

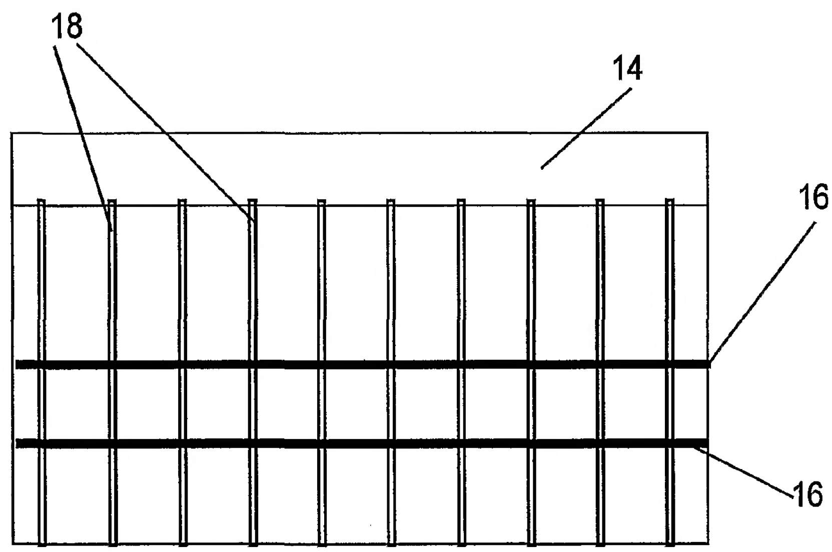


FIGURE 7

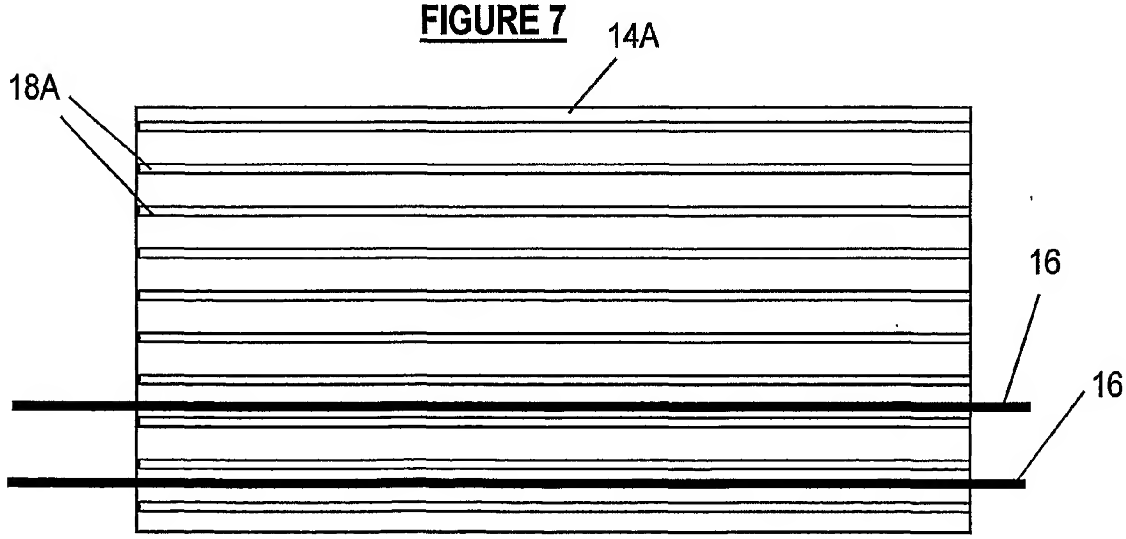


FIGURE 8

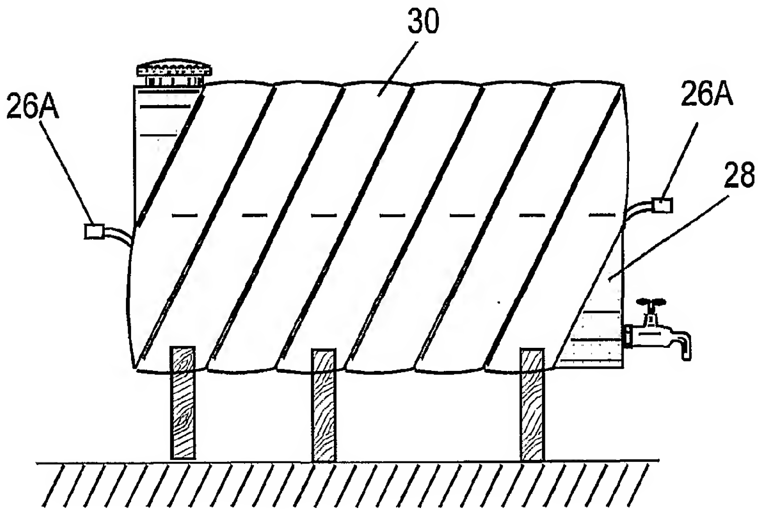


FIGURE 9

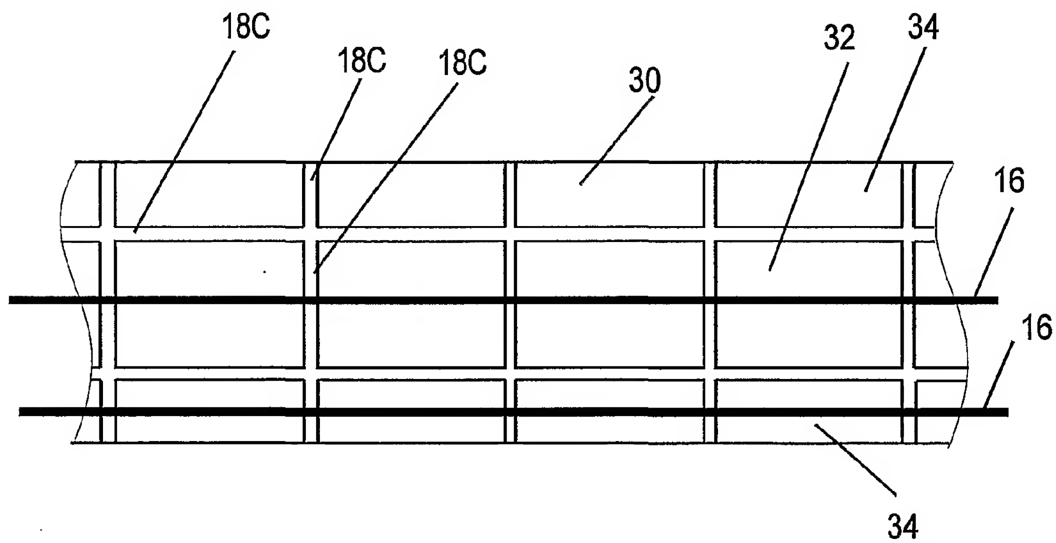


FIGURE 10

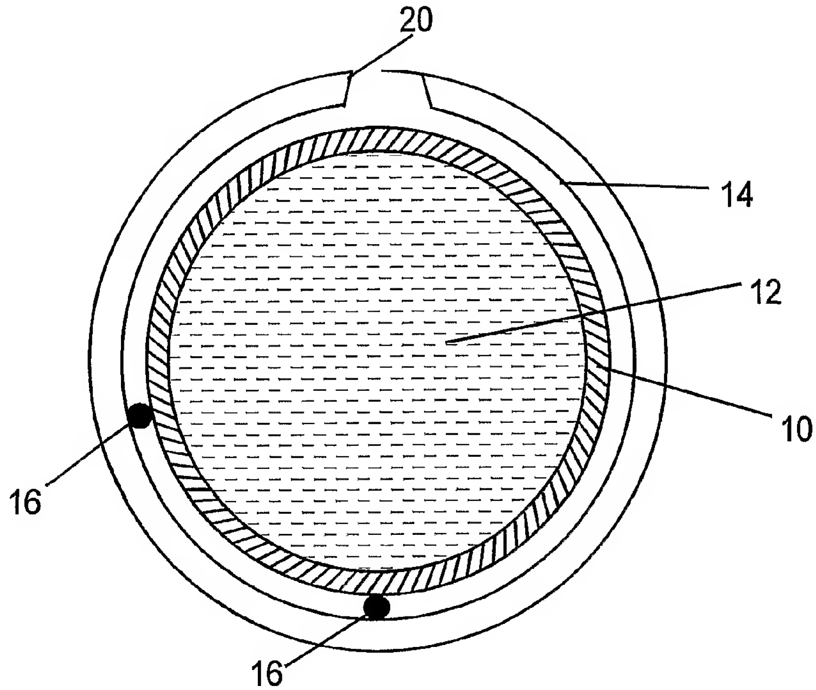


FIGURE 11

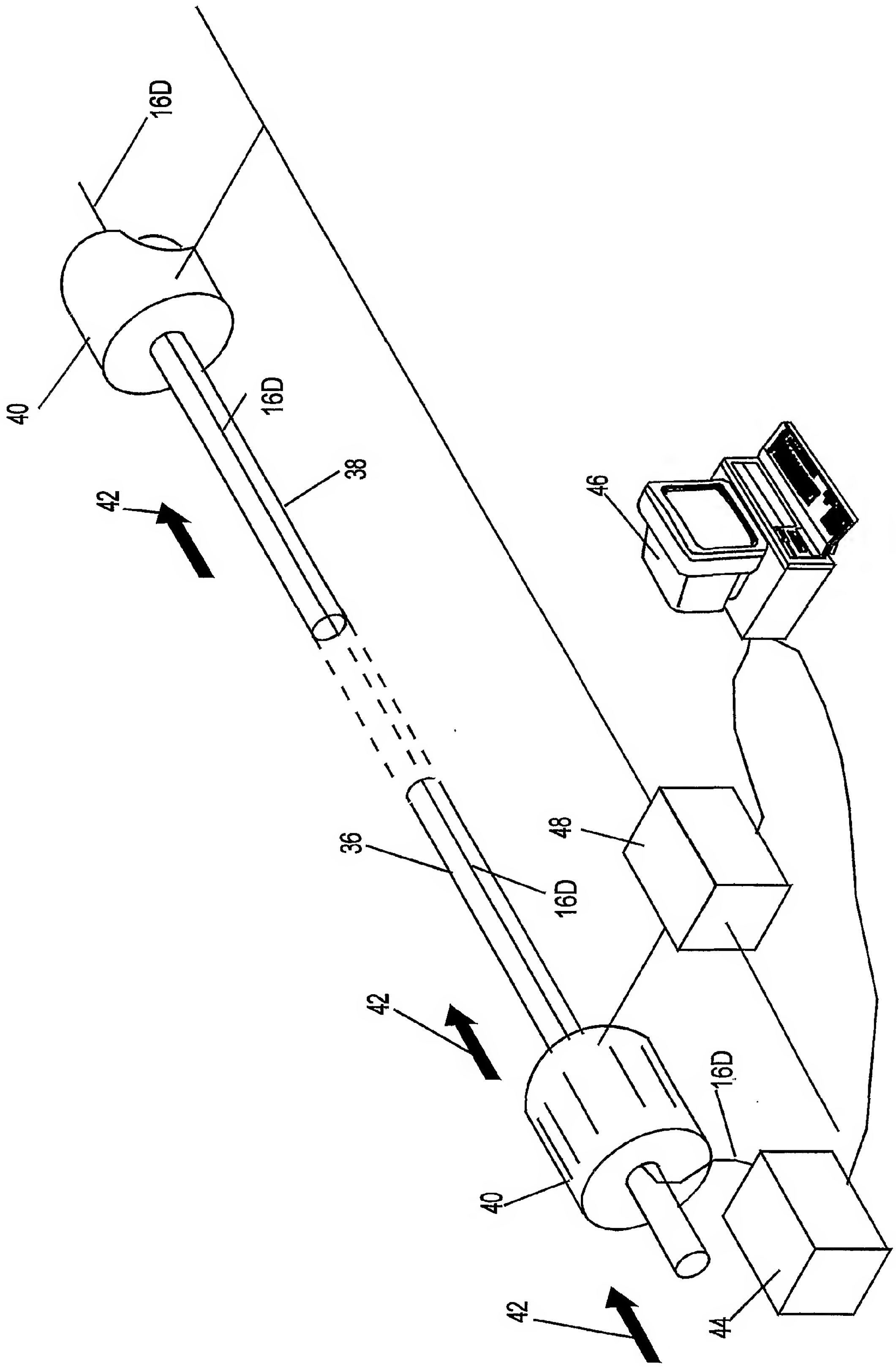


FIGURE 12

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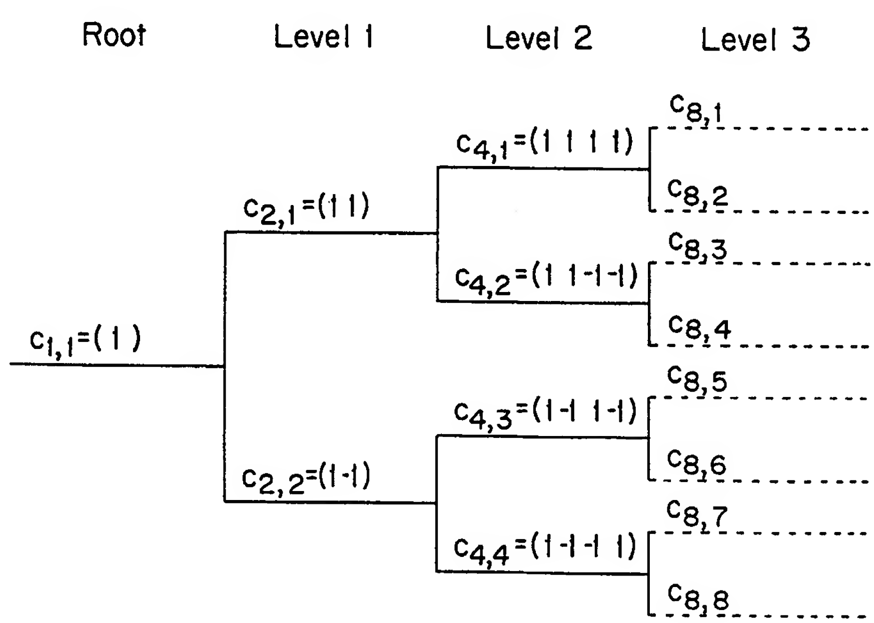

(57) Abstract

Variable spreading factors and multi-code transmissions are flexibly accomodated by assigning spreading codes in accordance with the described techniques. Spreading codes are assigned so that the control channel is orthogonal to all physical channels in the composite spread spectrum signal. Power balance between in-phase (I)and quadrature (Q) branches in the transmitter is also provided by assigning physical channels to appropriate branches and splitting physical channels, where necessary.

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CHANNELIZATION CODE ALLOCATION FOR RADIO COMMUNICATION SYSTEMS

BACKGROUND

This invention generally relates to variable data rate transmissions and, more particularly, to techniques for efficiently allocating spreading codes for variable rate data transmissions.

Cellular radio communication systems have recently been developed that use spread spectrum modulation and code division multiple access (CDMA) techniques. In a typical direct sequence CDMA system, an information data stream to be transmitted is superimposed on a much-higher-symbol-rate data stream sometimes known as a spreading sequence. Each symbol of the spreading sequence is commonly referred to as a chip. Each information signal is allocated a unique spreading code that is used to generate the spreading sequence typically by periodic repetition. The information signal and the spreading sequence are typically combined by multiplication in a process sometimes called coding or spreading the information signal. A plurality of spread information signals are transmitted as modulations of radio frequency carrier waves and are jointly received as a composite signal at a receiver. Each of the spread signals overlaps all of the other coded signals, as well as noise-related signals, in both frequency and time. By correlating the composite signal with one of the unique spreading sequences, the corresponding information signal can be isolated and decoded.

As radiocommunication becomes more widely accepted, it will be desirable to provide various types of radiocommunication services to meet consumer demand. For example, support for facsimile, e-mail, video, internet access, etc. via radiocommunication systems is envisioned. Moreover, it is expected that users may wish to access different types of services at the same time. For example, a video conference between two users would involve both speech and video support. Some of these different services will require relatively high data rates compared with speech service that has been conventionally supplied by radio communication systems, while other services will require variable data rate service. Thus, it is anticipated that future

radio communication systems will need to be able to support high data rate communications as well as variable data rate communications.

Several techniques have been developed to implement variable rate communications in CDMA radio communication systems. From the perspective of transmitting data at varying rates, these techniques include, for example, discontinuous transmission (DTX), variable spreading factors, multi-code transmission and variable forward error correction (FEC) coding. For systems employing DTX, transmission occurs only during a variable portion of each frame, i.e., a time period defined for transmitting a certain size block of data. The ratio between the portion of the frame used for transmission and the total frame time is commonly referred to as the duty cycle γ. For example, when transmitting at the highest possible rate, i.e., during the entire frame period, $\gamma = 1$, while for zero rate transmissions, e.g., during a pause in speech, $\gamma = 0$. DTX is used, for example, to provide variable data rate transmissions in systems designed in accordance with the U.S. standard entitled "Mobile Station-Base Station Compatibility Standard for Dual-Mode Wideband Spread Spectrum Cellular System", TIA/EIA Interim Standard TIA/EIA/IS-95 (July 1993) and its revision TIA/EIA Interim Standard TIA/EIA/IS-95-A (May 1995). Such standards that determine the features of U.S. cellular communication systems are promulgated by the Telecommunications Industry Association and the Electronic Industries Association located in Arlington, Virginia.

Varying the spreading factor is another known technique for providing variable data rate communication. As mentioned above, DS-CDMA spread spectrum systems spread data signals across the available bandwidth by multiplying each of the data signals with spreading sequences. By varying the number of chips per data symbol, i.e., the spreading factor, while keeping the chip rate fixed, the effective data rate can be controllably varied. In typical implementations of the variable spreading factor approach, the spreading factor is limited by the relationship to $SF = 2^k x SF_{min}$ where SF_{min} is the minimum allowed spreading factor corresponding to the highest allowed user rate.

Another known technique for varying the transmitted data rate is commonly referred to as multi-code transmission. According to this technique, data is transmitted using a variable number of spreading codes where the exact number of codes used depends on the instantaneous user bit rate. Effectively, this means allocating a variable number of physical channels to a connection to provide variable bandwidth. An example of multi-code transmission is described in U.S. Patent Application Serial No. 08/636,648 entitled "Multi-Code Compressed Mode DS-CDMA Systems and Methods", filed on April 23, 1996, the disclosure of which is incorporated here by reference.

Yet another technique for varying the transmitted data rate in radio communication systems involves varying the FEC. More specifically, the rate of the forward error correction (FEC) coding is varied by using code-puncturing and repetition or by switching between codes of different rates. In this way the user rate is varied while the channel bit rate is kept constant. Those skilled in the art will appreciate the similarities between varying the FEC and a variable spreading factor as mechanisms to implement variable rate transmission.

In both the uplink and downlink, it is desirable that any number of logical channels can be transmitted simultaneously to support a single connection between a base station and a mobile station to support various data rates. To transmit these logical channels over the radio interface, a number of physical channels are allocated. These physical channels are separated using different spreading codes (channelization codes), i.e., multicode transmission is used. Each physical channel can have one of several possible data rates, i.e., one of several possible spreading factors is used when spreading the data transmitted on the physical channel. To date, however, a flexible solution which allocates code words to physical channels taking into consideration the codes which have already been allocated to other channels and power considerations associated with the in-phase (I) and quadrature (Q) transmitter branches has not been provided.

Accordingly, it would be desirable to create new techniques and systems for allocating spreading codes in a flexible manner that supports multicode transmissions and variable spreading factors, and that optimizes power efficiency.

SUMMARY

These and other problems associated with previous communication systems are solved by Applicants' invention, wherein spreading codes are allocated for physical channels taking into consideration the spreading codes already allocated to other physical channels to be transmitted in parallel therewith. For example, if the physical channel being allocated a spreading code is a control channel (PCCH), then techniques according to the present invention investigate whether another physical channel on either the I or Q branches of the transmitter has already been assigned a spreading code so that the PCCH can be allocated a spreading code which makes the PCCH orthogonal to all other physical channels used in the composite spread spectrum signal. Moreover, for physical data channels (PDCH), techniques according to the present invention determine if any other channels have previously been assigned spreading codes on the same I or Q branch as the channel currently under investigation. If so, this PDCH is allocated a spreading code that makes the PDCH orthogonal to other PDCHs in the same branch, as well as to the PCCH.

According to other exemplary embodiments of the present invention, in addition to assigning a spreading code to each physical channel, the physical channels are also allocated between the I and Q branches of the transmitter in a manner intended to balance power between the two branches and improve power amplifier performance. For example, if the data rate associated with a connection to be set up is relatively low, then the connection may be supported by one PDCH and one PCCH, one of which is assigned to the I branch of the transmitter and the other to the Q branch. If, however, the data rate associated with a connection to be set up is relatively high, then assigning the PDCH to one branch and the PCCH to the other creates a large power discrepancy between the two branches. In such a case, the data can be transmitted on two PDCHs

each of which are allocated to the I and Q branches of the transmitter, respectively, and the control channel can be allocated to either the I or Q branch.

BRIEF DESCRIPTION OF THE DRAWINGS

The features and objects of Applicants' invention will be understood by reading this description in conjunction with the drawings, in which:

FIG. 1A is a block diagram representation of an exemplary transmitter structure in which the present invention can be implemented;

FIG. 1B illustrates an alternative scrambling technique which can be implemented in the transmitter of FIG. 1A;

FIG. 2 is an exemplary code tree;

FIG. 3 is a flowchart depicting allocation of physical channels between the I and Q branches of a transmitter according to an exemplary embodiment of the present invention; and

FIG. 4 is a flowchart illustrating the allocation of spreading codes to physical channels according to the present invention.

DETAILED DESCRIPTION

While this description is written in the context of cellular communications systems involving portable or mobile radio telephones, it will be understood by those skilled in the art that Applicants' invention may be applied to other communications applications.

According to exemplary embodiments of the present invention, CDMA systems can support variable bit rate services, such as speech, by providing control information in each frame which specifies the instantaneous data symbol rate for that frame. In order to accomplish this in a regular time interval, physical channels can be organized in frames of equal length (timewise). Each frame carries an integer number of chips and an integer number of information bits.

Using this exemplary frame structure, bit rate control information can be provided for every CDMA frame by transmitting this information on a separate

physical channel. The physical channels carrying the data and the control information (e.g., including pilot/reference symbols for channel estimation, power control commands and rate information of the data) can be denoted as physical data channel (PDCH) and physical control channel (PCCH), respectively. Each connection between a mobile station and a base station will be supported by a PCCH and at least one PDCH. The spreading code, symbol rate, or equivalently spreading factor, of the PCCH are known a priori to the receiver. In this way, the receiver can determine the data rate of the PDCH(s) from the PCCH prior to demodulating/decoding the PDCH(s). Exemplary techniques for handling BRI information are described in commonly-assigned, copending U.S. Patent Application Serial No. ________________, entitled "Low-Delay Rate Detection for Variable Rate Communication Systems" to Dahlman et al., filed on an even date herewith.

Many potential advantages are attributable to variable rate transmission. For example, interference can be reduced for various users of the system since the chip rate is kept constant and a lower bit rate gives a higher spreading factor, thus allowing a lower transmit power. Those skilled in the art will readily appreciate how this ability to vary the information rate in a CDMA system can be used advantageously to vary other parameters. However, techniques for efficiently allocating spreading codes to the various physical channels (i.e., PCCH and PDCH(s)) are needed and described below.

A physical channel is a bit stream of a certain rate, that is spread using a certain code and allocated to either the in-phase (I) or quadrature (Q) branch in a transmitter. Variable rate services are supported through spreading with a variable spreading factor as described above. A number of data streams are spread to the chip rate using Walsh codes of different length, followed by summation and, if desired, scrambling. The structure of an exemplary transmitter (usable, e.g., in either a base station or a mobile station) which performs these spreading, summing and scrambling operations is illustrated in Figure 1A.

Therein, a first data stream I_1 is supplied to multiplier 10 having a data rate of R_1 which is equal to the chip rate R_c divided by the spreading factor SF_{I1} for that data stream. This data stream is spread with a channelization code word C_{I1} having a length

of 2^k chips which is selected such that the output of multiplier 30 has a chip rate R_c by selecting a value for k that is related to the desired data rate of physical channel I_1. For example, a physical channel data rate of 250 kbps is spread to a chip rate of 4 Mcps by using a channelization code of 16 (2^4) chips long. More details regarding the allocation of a particular channelization code according to the present invention are described below. Similarly, additional data streams are supplied to multipliers 12, 14 and 16 (and other branches which are unillustrated) to spread their respective data streams with channelization code words having a length which is selected to result in a chip rate R_c. The rate of the data streams can be limited to such an interval that the spreading factors used are larger or equal to a predetermined SF_{min}. Each physical channel is then weighted by respective amplifiers 18, 20, 22 and 24. The weights can be individually chosen to allocate power to each physical channel so that predetermined quality requirements, e.g., the bit error rate of each physical channel, are satisfied. The physical channels in the "I" branch of the transmitter are summed at summer 26. Similarly, the physical channels in the "Q" branch of the transmitter are summed at summer 28. Scrambling, if desired, is then performed on the superimposed physical channels. This can be done in at least two ways. First, as shown in Figure 1A, scrambling can be performed by forming the I and Q pairs as a complex number at blocks 30 and 32 and then multiplying the result with another complex number (i.e., the complex valued scrambling code $c_{scramb}=c_I+jc_Q$) at block 34. Scrambling can also be performed on the I and Q branches separately as illustrated in Figure 1B, by multiplying I and Q with two real valued scrambling codes c_I and c_Q at blocks 36 and 38. The scrambling code is clocked at the chip rate. The resultant signal is output, e.g., to transmit signal processing circuitry (e.g. a QPSK or O-QPSK modulator followed by, possibly, pulse-shaping filters), amplified by a transmit power amplifier (not shown) and ultimately coupled to an antenna (also not shown).

The Walsh codes used for spreading at multipliers 10-16 can be viewed in a tree like manner, as illustrated in Figure 2. Codes on the same level in the tree are orthogonal and have the same spreading factor. Thus, codes $c_{4,1}$, $c_{4,2}$, $c_{4,3}$ and $c_{4,4}$ are orthogonal codes each of which have the same spreading factor, i.e., the same number

of chips. If a physical channel is spread with a first code in the tree, and another physical channel is spread with another code which is (1) not the same as the first code, (2) not to the left of the first code on the path to the root of the tree and (3) not in the subtree which has the first code as the root, then the two spread physical channels will be orthogonal. For example, if the PCCH is allocated code $c_{4,1}$ and a PDCH is allocated code $c_{8,5}$, then these two spread channels would be orthogonal. If, however, the PDCH was allocated code $c_{8,1}$ or $c_{8,2}$, then the PCCH and PDCH would be non-orthogonal. Every physical channel is allocated a spreading code from the tree, with spreading factors matching the respective data rates. As the data rate varies for a particular PDCH, a code from a different level of the tree will be allocated. For example, increasing data rates will cause code selection to move to the left in the tree, while for decreasing data rates code selection will move to the right. Thus, a typical variable rate PDCH will typically move up and down along a certain path in the code tree as its data rate varies. Allocation of physical channels to the I and Q branches of the transmitter, as well as codes from the code tree in Figure 2 as spreading codes (e.g., c_{I1}, c_{Q1}, etc. in Figure 1A) can be made according to the following rules in accordance with the present invention.

Figure 3 is a flowchart which illustrates an exemplary technique for allocating the physical channels between the I and Q branches of a transmitter according the present invention for the case where a single PDCH can be used (i.e., has sufficient bandwidth) to support a connection. Those skilled in the art will appreciate that this technique provides for a relatively balanced transmit power for the each of the I and Q branches which in turn provides better power amplifier performance. The flow begins at block 40 wherein it is determined whether the power that would be needed to transmit the single PDCH is significantly greater than that needed to transmit the PCCH. For example, if the PDCH is to be transmitted at a much higher rate than the PCCH or if the quality of service (QoS) requirements for the PDCH are higher, then the power requirements will be correspondingly higher. In such a case, the flow proceeds to block 42 wherein the data stream is split into two lower rate PDCHs. The three physical channels can then be allocated, for example, as illustrated in block 42 to

the I and Q branches in a manner which will help to more evenly balance the transmit power between these two branches. If, on the other hand, it is determined at block 40 that the PDCH is not to be transmitted at a significantly greater power than the PCCH, then the flow proceeds to block 44 wherein the control channel is allocated to one of the branches and the data channel to the other. Note that the particular selection of Q and I in blocks 42 and 44 is exemplary only and that these designations could of course be reversed.

Having assigned the physical channels to a respective one of the I and Q branches in the transmitter, the next allocation to be made according to the present invention is the selection of a spreading code for each of the physical channels. According to the present invention, the spreading code selected to spread the PCCH should be such that the PCCH is orthogonal to all of the other physical channels to be transmitted in the composite spread spectrum signal, i.e., orthogonal to all channels in both the I and Q branches. This characteristic is desirable because the PCCH must first be demodulated and decoded at the receiver to provide channel estimates which are used to process the data channels transmitted in the same spread spectrum signal. Accordingly, an exemplary technique for allocating spreading codes according to the present invention will now be described with respect to the flowchart of Figure 4. The flow begins at block 52 wherein it is determined whether the present channel that is being allocated a spreading code is a data channel or a control channel. If the channel currently being allocated a spreading code is a PDCH then the flow proceeds to block 54. Therein, this PDCH is allocated a spreading code which makes the PDCH orthogonal to the PCCH (if the PCCH has already been allocated a spreading code) and which makes the PDCH orthogonal to any other PDCH that is on the same I or Q branch of the transmitter. For example, suppose that at the time this particular PDCH is being allocated a spreading code that the PCCH has already been allocated code $c_{4,1}$ and another PDCH has already been allocated code $c_{8,5}$. Further, assume that this particular PDCH is to be transmitted at a data rate that requires a level 3 code with respect to the code tree of Figure 2. According to the present invention, this particular PDCH could then be allocated any of codes $c_{8,3}$, $c_{8,4}$, $c_{8,6}$, $c_{8,7}$ and $c_{8,8}$. This PDCH

could not be allocated to codes $c_{8,1}$ or $c_{8,2}$ since such allocations would result in non-orthogonality with the control channel. This PDCH could, however, be allocated code $c_{8,5}$ if it is assigned to the opposite transmitter branch of the PDCH which has already been assigned this spreading code.

The flow then proceeds to block 56 whereupon more codes are allocated if additional channels remain. Otherwise the process terminates. If, at block 52, a control channel is being evaluated for spreading code allocation, then the flow proceeds to block 58. Therein, a code is selected which makes the control channel orthogonal to all channels previously allocated codes so that the PCCH can be readily decoded and demodulated at the receiver to provide channel estimates for use and evaluating the data channels.

It will be understood that Applicants' invention is not limited to the particular embodiments described above and that modifications may be made by persons skilled in the art. The scope of Applicants' invention is determined by the following claims, and any and all modifications that fall within that scope are intended to be included therein.

We Claim:

1. A transmitter having an in-phase (I) branch and a quadrature (Q) branch for transmitting a composite, spread spectrum signal including at least two physical channels, said transmitter comprising:

means, associated with said I branch, for spreading data associated with one of said at least two physical channels using a first spreading code to generate a first spread physical channel; and

means, associated with said Q branch, for spreading data associated with another of said at least two physical channels using a second spreading code to generate a second spread physical channel;

wherein said first and second spreading codes have a different number of chips and said first and second spreading codes are selected so that said first and second spread physical channels are orthogonal to one another.

2. The transmitter of claim 1, wherein said one of said at least two physical channels is a control channel (PCCH) and said another of said at least two physical channels is a data channel (PDCH).

3. The transmitter of claim 2, further comprising:

means for balancing power associated with said I and Q branches of said transmitter by selectively allocating said at least two physical channels to said I and Q branches based on transmit power requirements.

4. The transmitter of claim 3, wherein said at least two physical channels include a second PDCH which is spread using a third code to generate a third spread physical channel, and wherein said means for balancing power allocates said second PDCH to a same branch of said transmitter as said PCCH to based on said transmit power requirement.

5. The transmitter of claim 4, wherein said second and third spread physical channels are orthogonal.

6. The transmitter of claim 4, wherein said second and third spread physical channels are non-orthogonal.

7. The transmitter of claim 4, wherein said second and third codes are the same codes.

8. A method for allocating spreading codes to a plurality of physical channels to be transmitted in a composite spread spectrum signal in a radio communication system comprising the steps of:

allocating a first spreading code having a first number of chips to a control channel so that said control channel is orthogonal to others of said plurality of physical channels in said composite spread spectrum signal; and

allocating a second spreading code having a second number of chips different from said first number of chips to a first data channel, which second spreading code is selected such that said control channel and said first data channel are orthogonal to one another.

9. The method of claim 8, wherein said control channel conveys reference information usable to make channel estimates.

10. The method of claim 8 further comprising the step of:

allocating a third spreading code having a third bit length to a second data channel, said third spreading code selected such that said control channel and said second data channel are orthogonal to one another.

11. The method of claim 10, wherein said first and second data channels are orthogonal.

12. The method of claim 10, wherein said first and second data channels are non-orthogonal.

13. The method of claim 10, wherein said second and third spreading codes are the same codes.

14. The method of claim 10, further comprising the steps of:

assigning said second data channel to one of an I and a Q branch in a transmitter; and

assigning said third data channel to the other of said I and Q branches.

15. The transmitter of claim 1, further comprising:

means for scrambling said first and second spread physical channels of said I and Q branches,

16. A method for allocating spreading codes to a plurality of physical channels to be transmitted in a composite spread spectrum signal in a radio communication system comprising the steps of:

allocating a first spreading code having a first number of chips to a first data channel; and

allocating a second spreading code having a second number of chips different from said first number of chips to a control data channel, which second spreading code is selected such that said control channel and said first data channel are orthogonal to one another.

17. The method of claim 16, wherein said control channel conveys reference information usable to make channel estimates.

18. The method of claim 16 further comprising the step of:

allocating a third spreading code having a third bit length to a second data channel, said third spreading code selected such that said control channel and said second data channel are orthogonal to one another.

19. The method of claim 18, wherein said first and second data channels are orthogonal.

20. The method of claim 18, wherein said first and second data channels are non-orthogonal.

21. The method of claim 18, wherein said second and third spreading codes are the same codes.

22. The method of claim 18, further comprising the steps of:

assigning said second data channel to one of an I and a Q branch in a transmitter; and

assigning said third data channel to the other of said I and Q branches.

Fig. 1A

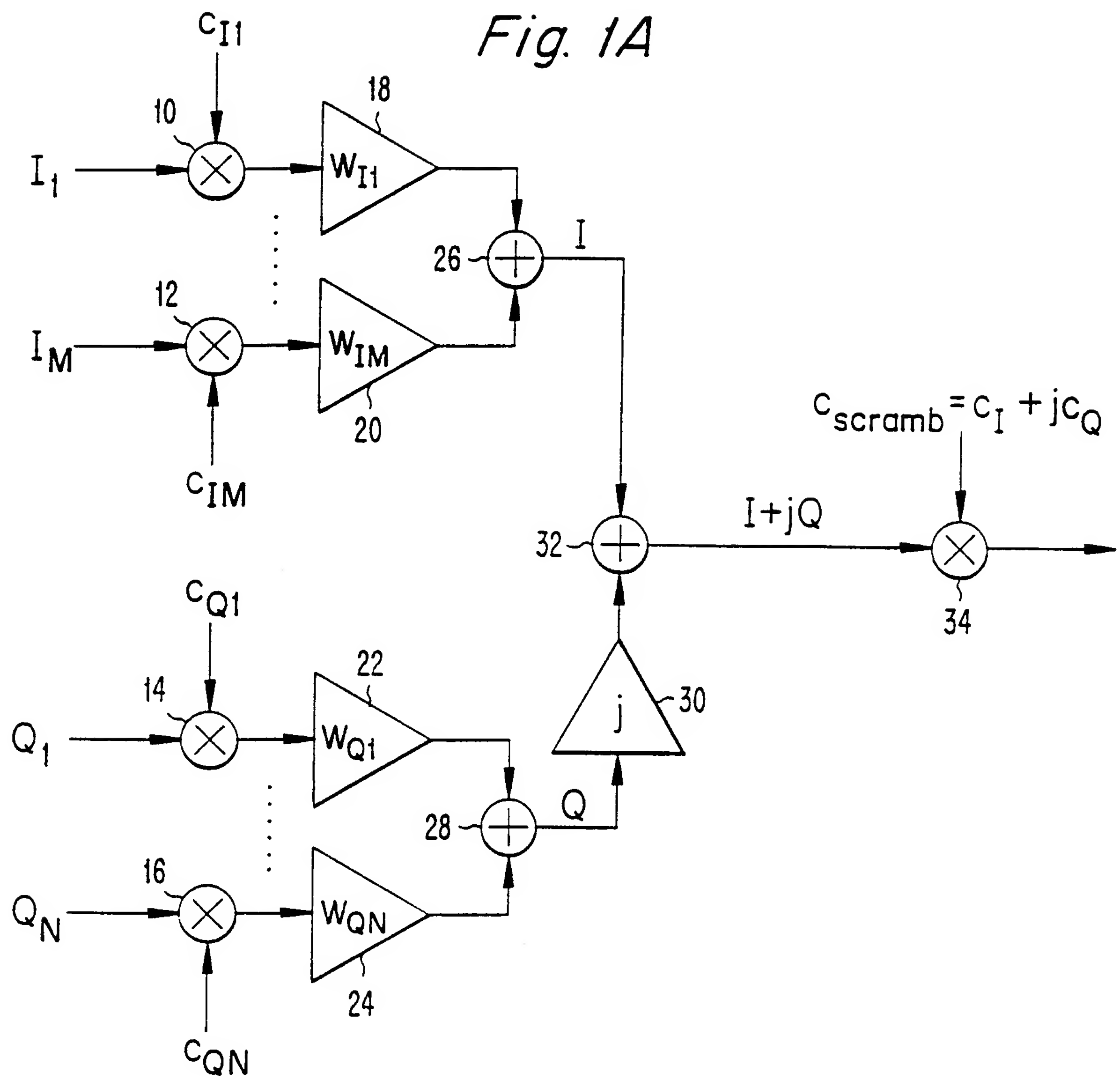

c_{I1}
10
18
I_1
W_{I1}
26
I
12
I_M
W_{IM}
20
c_{IM}
$c_{scramb} = c_I + jc_Q$
32
I+jQ
34
c_{Q1}
22
14
Q_1
W_{Q1}
j
30
28
Q
16
Q_N
W_{QN}
24
c_{QN}

Fig. 1B

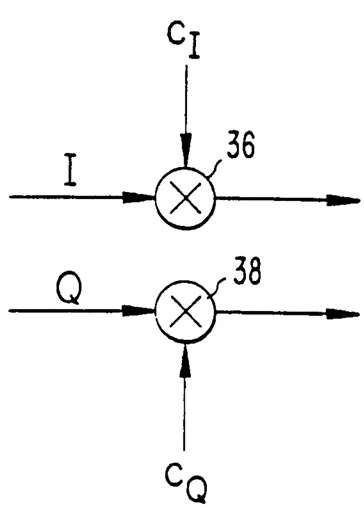

c_I
36
I
Q
38
c_Q

Fig. 2

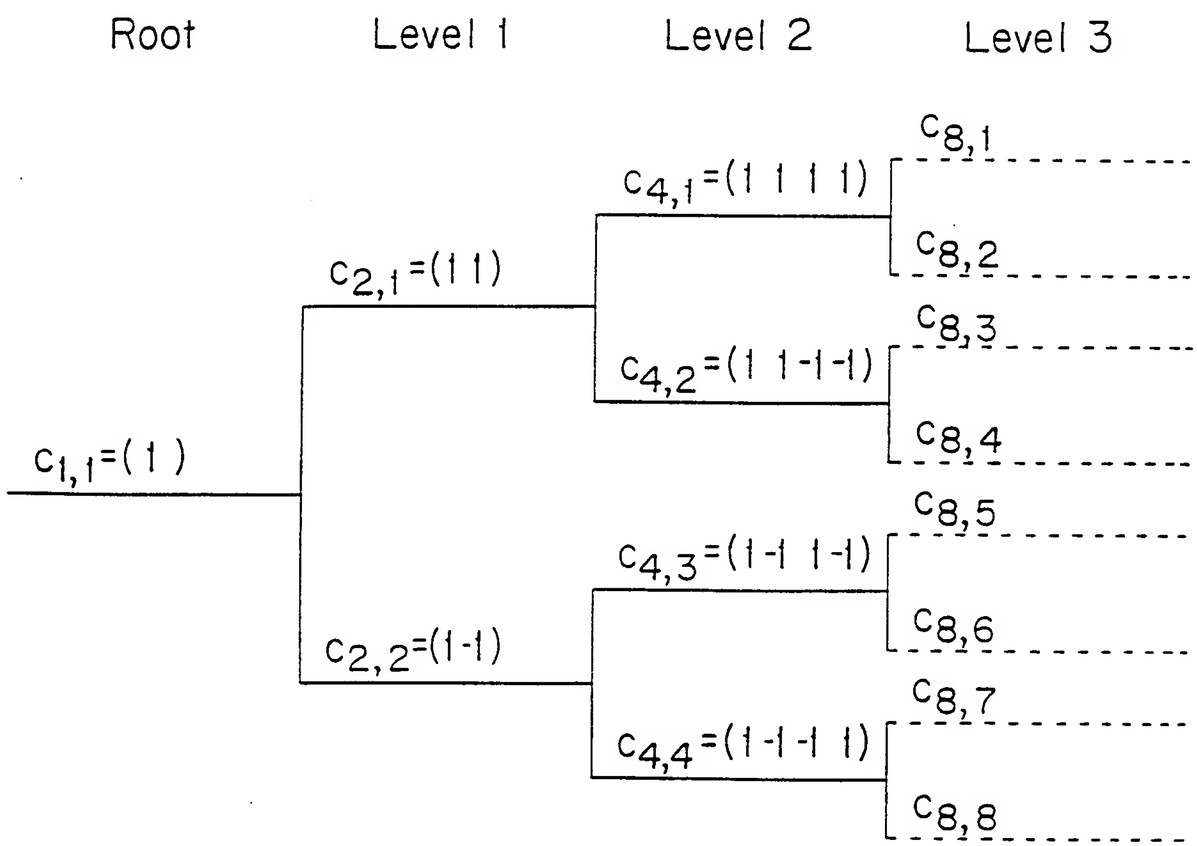
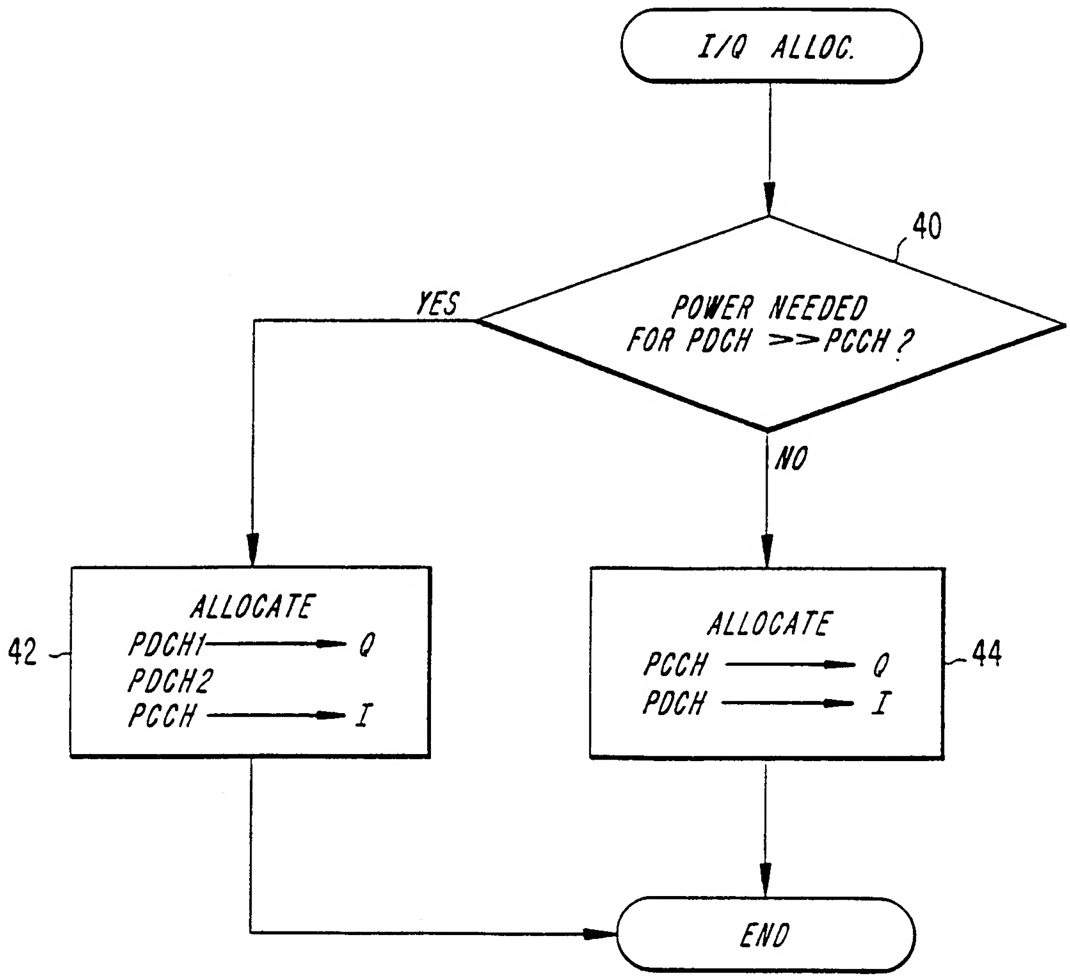

Root
Level 1
Level 2
Level 3
$c_{1,1}=(1)$
$c_{2,1}=(1\ 1)$
$c_{2,2}=(1\ -1)$
$c_{4,1}=(1\ 1\ 1\ 1)$
$c_{4,2}=(1\ 1\ -1\ -1)$
$c_{4,3}=(1\ -1\ 1\ -1)$
$c_{4,4}=(1\ -1\ -1\ 1)$
$c_{8,1}$
$c_{8,2}$
$c_{8,3}$
$c_{8,4}$
$c_{8,5}$
$c_{8,6}$
$c_{8,7}$
$c_{8,8}$

Fig. 3


I/Q ALLOC.
40
POWER NEEDED FOR PDCH >>PCCH ?
YES
NO
42
ALLOCATE
PDCH1 → Q
PDCH2
PCCH → I
ALLOCATE
PCCH → Q
PDCH → I
44
END

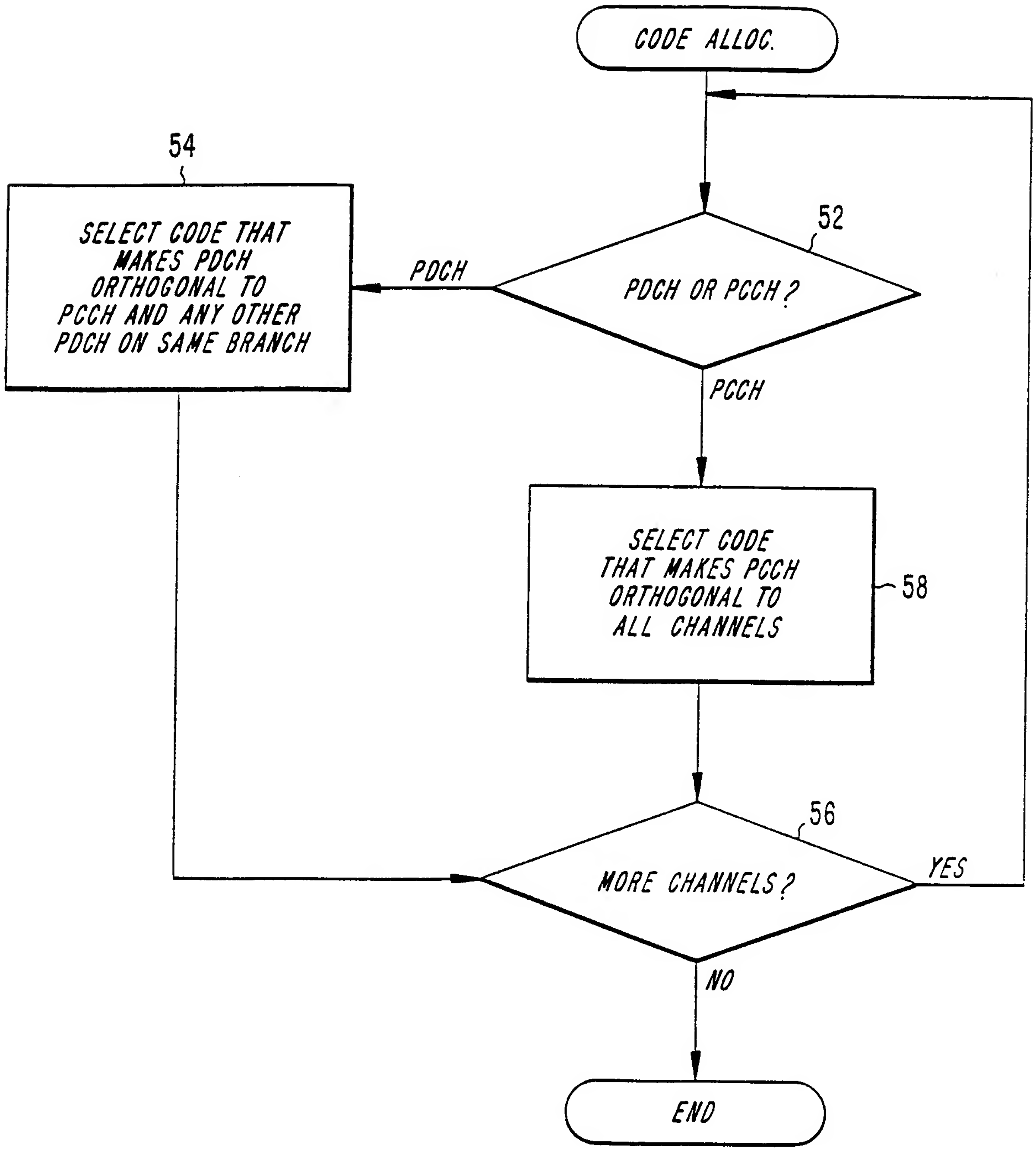

CODE ALLOC.
54
SELECT CODE THAT MAKES PDCH ORTHOGONAL TO PCCH AND ANY OTHER PDCH ON SAME BRANCH
PDCH
52
PDCH OR PCCH?
PCCH
SELECT CODE THAT MAKES PCCH ORTHOGONAL TO ALL CHANNELS
58
56
MORE CHANNELS?
YES
NO
END

Fig. 4

INTERNATIONAL SEARCH REPORT

Int ional Application No
PCT/SE 98/01317

A. CLASSIFICATION OF SUBJECT MATTER
IPC 6 H04J13/04 H04B7/26

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 6 H04J H04B

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US 5 559 788 A (ZSCHEILE JR JOHN W ET AL) 24 September 1996 see abstract see column 2, line 8 - line 20 see column 3, line 53 - column 4, line 12; figure 2 see column 5, line 33 - line 50; figures 6-8 ---	1,8,16
A	WO 95 03652 A (QUALCOMM INC) 2 February 1995 see abstract see page 4, line 3 - line 18 see page 14, line 32 - page 16, line 32; figure 2; table I --- -/--	1,8,16

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

° Special categories of cited documents :

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"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
21 October 1998	26/10/1998

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL - 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Bossen, M

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INTERNATIONAL SEARCH REPORT

Int ional Application No
PCT/SE 98/01317

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
A	US 5 544 156 A (TEDER PAUL M ET AL) 6 August 1996 see column 3, line 49 - column 4, line 19 -----	2,4,8,16

1

Form PCT/ISA/210 (continuation of second sheet) (July 1992)

INTERNATIONAL SEARCH REPORT

Information on patent family members

International Application No
PCT/SE 98/01317

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WO 9503652 A	02-02-1995	AU 7368294 A	20-02-1995
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		WO 9530289 A	09-11-1995